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PROBLEMS OF SPACE RADIO COMMUNICATION

by N. T. Petrovich and Ye. F. Kamnev

Izd-vo Sovetskoye Radio,
Moscow, 1965

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Translation of "Voprosy kosmicheskoy radiosvyazi."
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Achievement of

V. M. Komarov, K. P. Feoktistov and B. B. Yegorov

This book considers two basic aspects of space radio communication: use of artificial Earth satellites (AES) and the Moon for the creation of global and local ground communication systems of various types, and the provision of two-way radio communication between the Earth and spacecraft.

The general principles of the design of communication systems using satellite relay are presented and examples are given of global and local ground communication systems. On the basis of the analysis of the quantities appearing in the active and passive communication equations, an evaluation is made of the optimal operating frequencies and the power requirements for active and passive relay.

A power requirement analysis is made for space radio communication links within the limits of the solar system.

A review is presented of the published data on systems for spacecraft communication issued in the USSR and USA, as well as data on satellite communication systems.

The book is intended for students at the college-level radio institutes and for a broad spectrum of specialists who are to some degree concerned with space radio communications.

INTRODUCTION

The problem of space radio communication has two primary aspects: the first involves the use of artificial Earth satellites (AES) and the various cosmic bodies for the purpose of creating global and local systems for Earth radio communication in various forms--telegraphy, telephony and television, while the second encompasses the questions of providing radio communication for spacecraft, both in the near-Earth space and at considerable distances from Earth.

In spite of the large number of publications on this problem which have appeared in recent years in both the Soviet and foreign press, at present there is no text available which presents a sequential development of the technique of the design of space radio links with specific examples of such calculations.

The authors hope to fill this gap to some degree and to present the methodology of the engineering design of space radio links, making extensive use of published material.

Here we shall consider two aspects of this problem: the use of AES and the Moon as active and passive radio relay stations in ground systems communication, and the provision for radio communication with spacecraft traveling in the vicinity of the planets of the solar system.

This book is the record of lectures on several sections (space radio /6
communication, the problems of improving noise immunity and the rate of transmission of information over communication channels, etc.) of the general course "Fundamentals of Radio Communication," presented at the All-Union Correspondence Electrotechnical Institute of Communications by N. T. Petrovich in 1962-64. The program and the content of these sections of the course were prepared jointly by the authors.

The authors wish to express their sincere thanks to Dr. Tech. Sci. Prof. V. A. Smirnov for his valuable comments and recommendations made during his review of the manuscript. In addition, the authors are very grateful to T. I. Zelenoborskaya for extensive assistance in the preparation of the manuscript.

CONTENTS

CHAPTER I. USE OF ARTIFICIAL EARTH SATELLITES AS RADIO RELAY STATIONS IN GROUND COMMUNICATION SYSTEMS

1. The Promise of the Use of AES as Radio Relay Stations /7*

The existing systems for long-range communication have several very serious deficiencies. We shall list the foremost of these.

1. The long and short radio wavebands (10-100 kcps, 1.5-30 Mcps), over which it is possible in principle to organize a global radio communication system without intermediate retransmitting stations, have limited frequency spectra. For the same reason both of these bands have comparatively low handling capacity and are severely overloaded at the present time. Some idea of the saturation of these bands is given in the article of S. I. Nadenenko (ref. 1), indicating the utilization of these bands during the ten-year period from 1930 to 1940.

Figure I-1 shows graphs of the growth of the number of channels (solid line) and the number of wavelengths used (dashed line) during the period from 1930 to 1940 in the 10-100 kcps and 1.5-30 Mcps bands.

In 1930 there were 6,203 officially registered radio stations in the range from 10 to 200 m (1.5-30 Mcps). In 1940 the number of officially registered stations in this range, not counting amateur and military stations, reached /9 28,407. By 1948 the officially registered requests for operating frequencies in this band exceeded 200,000 (ref. 2). At the present time the overloading of the HF band has become even greater. The existence of such a large number of operating stations has led to the situation where the primary form of interference at the short wavelengths is the noise from neighboring stations.

The limitation of the frequency spectrum in the region of the long and short wavelengths and the propagation characteristics of these bands (presence of pronounced dispersion) prevent the transmission of wide-band information, particularly broadcast television.

Another deficiency of the HF communication systems is the great dependence of their operation on the conditions of the atmosphere and, as a result of this, the susceptibility of the radio communication to failure. Radio links which pass through the polar absorption zone or lie in this zone are most frequently subjected to impairment of communication, including complete blockage for several days during the time of severe ionospheric storms (refs. 3, 4, 5 and 6).

*Numbers given in the margin indicate the pagination in the original foreign text.

2. The radio communication systems using the diffuse propagation of the radio waves in the UHF band (tropospheric and ionospheric scatter radio links), which have received major development in recent years as a result of the fact that they are practically immune to the effect of the ambient conditions, do not provide sufficiently high quality of transmission, have a limited passband (also unsuitable for television transmission) and relatively high cost. In addition, these communication systems cannot be used on many transoceanic lines as a result of the impossibility of installation of intermediate retransmitting points.

3. The underwater transoceanic cable lines also have a limited width of the transmission band, and their capacity cannot provide for the expected increase of radio communication on the transoceanic lines, which, as estimated by the U.S. specialists (ref. 3), by 1970 will increase by more than a factor of 7 in the comparison with 1950.

4. The waveguide communication lines using the millimeter band, providing an exceptionally wide passband, at the present time cannot be realized in view of their high cost. In addition, it is practically impossible to use such systems on the transoceanic lines (ref. 125).

The UHF and decimetric wavelength radio relay links with retransmitting stations on board AES do not have the disadvantages of the existing communication equipment; thus they seem to have great promise for the creation of /10 reliable and high-speed Earth communication systems.

Although indications had been made of the possibility of using AES in systems for long-range radio communication, including the transmission of television over long distances, long ago (refs. 7, 8 and 9), up to 1957 their use for the solution of the problem of long-range communication was considered to be scientific fantasy. The launch of the first Soviet Sputnik on 4 October 1957 opened up a new era in the mastery of outer space. The possibility of the creation of an artificial planet was proved in practice. This also proved the possibility of the use of AES as potential carriers for the rebroadcasting stations. The rapid development of rocket technology since 1957 gives a basis for presuming that already at the present stage the systems for ground communication using AES are advantageous, not only from the point of view of technical criteria, but also from the point of view of economics. The data on the weights of some Soviet satellites and space rockets give an impression of the rate of development of rocket technology (ref. 10):

First artificial satellite (4 October 1957)	83.6 kg
Second artificial satellite (3 November 1957)	508.3 kg
Third artificial satellite (15 May 1958)	1,327 kg
First space rocket (2 January 1959)	1,472 kg
Second space rocket (12 September 1959)	1,511 kg
Third space rocket (4 October 1959)	1,553 kg
First spacecraft satellite (15 May 1960)	4,540 kg
Second spacecraft satellite (19 August 1960)	4,600 kg

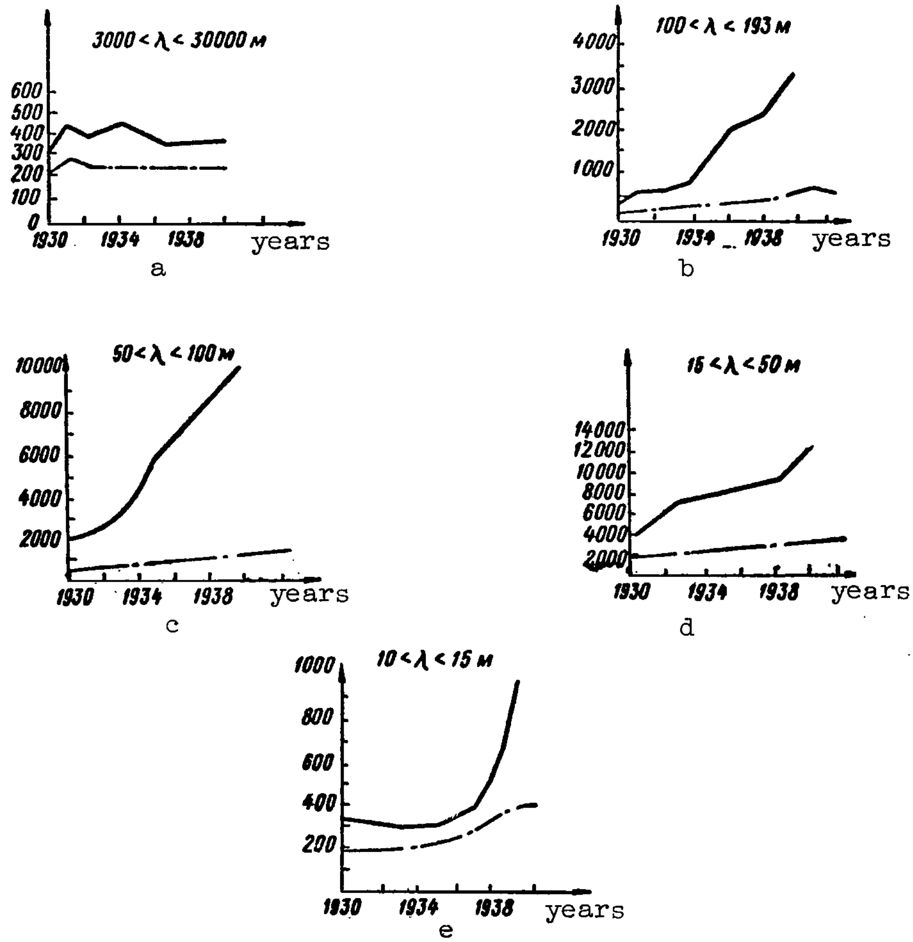


Figure I-1. Growth of number of channels and number of operational wavelengths λ in various bands from 1930 to 1940.

The increase of the payload and reliability of the rockets leads to a sharp reduction of the cost of communication systems using AES. Studies of U.S. authors (ref. 11) have shown that while in 1958 the cost per kg of launching a satellite into orbit around the Earth amounted to 2 million dollars, by 1964 the cost was about 200 dollars per kg.

After these introductory remarks let us turn to a more detailed consideration of the specific question of the design of a radio relay system for ground communication using retransmitting stations located on board AES. /11

2. Types of Orbit

If undisturbed motion is considered, then the Keplerian laws for the motion of planets around the Sun are applicable to the satellites orbiting about Earth (refs. 12 and 122).

In accordance with Kepler's first law one of the foci of the ellipse along which the AES moves must be located at the center of Earth; the second focus will be located at that same distance from the apogee of the satellite orbit as the center of Earth is from its perigee (fig. I-2).

According to Kepler's second law (law of areas) the radius-vector of the satellite describes equal areas in equal intervals of time (fig. I-3).

According to Kepler's third law the square of the time of revolution of the satellites around Earth is proportional to the cube of the semimajor axes of their orbits (fig. I-2)

$$T_c^2 = ka^3, \tag{I-1}$$

where T_c is the period of revolution of the satellite in orbit;

a is the magnitude of the semimajor axis of the ellipse;
k is a constant.

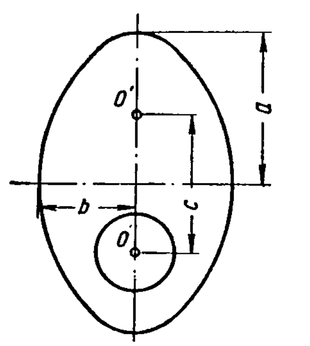


Figure I-2. Geometric illustration on Kepler's first law.

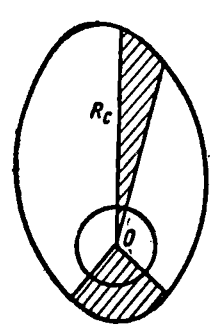


Figure I-3. Geometric illustration on Kepler's second law.

The exact form of the orbit is determined by the eccentricity e, equal to the ratio of the distance between the foci c to the semimajor axis a. For /12 e = 0 the ellipse degenerates into a circle, and the orbit will be circular. The circular and elliptic orbits are divided, in turn, into:

(1) polar, if the plane of the orbit includes the axis of rotation of the Earth;

(2) inclined, if the plane of the orbit does not include the axis of rotation of Earth and is not orthogonal to it;

(3) equatorial, if the orbital plane coincides with the equatorial plane of Earth (orthogonal to the axis of rotation of the Earth). In particular, the equatorial circular orbit at an altitude of h = 36,000 km above the surface of Earth is termed a stationary orbit. The period of revolution of a satellite in this orbit is 24 hours, and the satellite appears stationary for the observer on Earth.

3. Principles of Use of AES for Ground Communication Systems

AES can be used as active or passive signal relays for signals transmitted over long distances. Let us consider in more detail each of these cases.

Active Relay of Signals by AES. Two different methods of accomplishing the active relay of radio signals through the AES are possible (refs. 121 and 123):

(1) the method of immediate relay without delaying the information (conventional radio relay communication);

(2) the method of relay with delay (storage) of the signal on board the satellite.

In the first case the AES radio relay operates on the principle of the conventional radio relay station, i.e., it transmits the information to Earth directly after reception, without time delay. In this case, communication is possible only when the AES is simultaneously visible from both ground stations (fig. I-4).

In the second case, the operation proceeds as follows: during passage of the AES over the ground transmitting station the information to be transmitted is beamed to the AES and is there recorded on either magnetic tape, magnetic drum or magnetic disk. During the passage of the satellite over the ground /13 receiving station, the stored information is transmitted to Earth with the aid of a programmer or on interrogation from Earth (fig. I-5).

Relay stations with time delay of the information can accomplish /14 selective communication in definite geographical zones. It is possible to store a large volume of information and transmit it to points located thousands of kilometers from the transmitting station at any point on Earth. A global communication system of this type does not require high power or a large number of

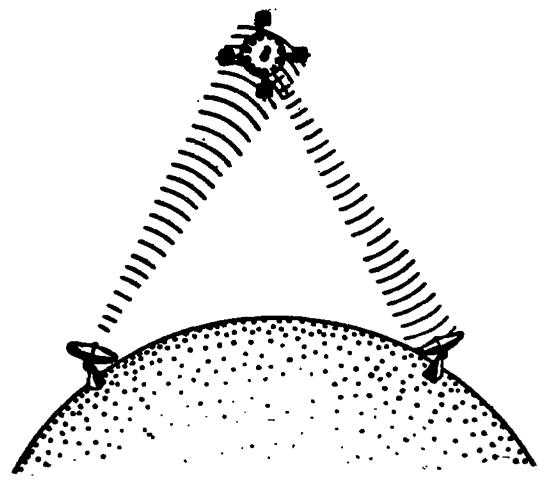


Figure I-4. Active relay.

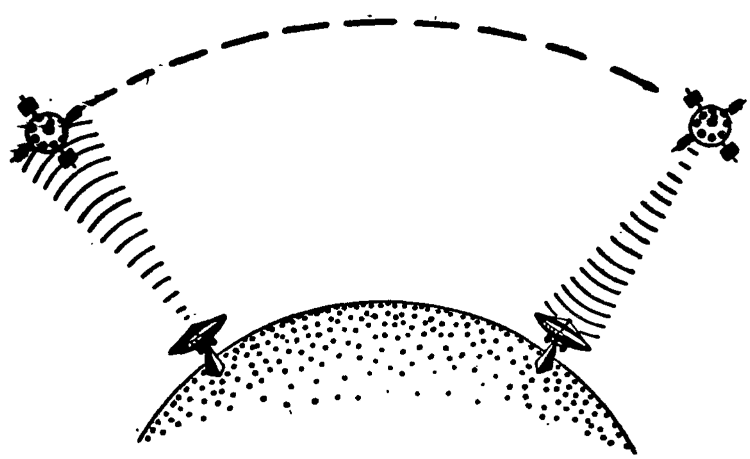


Figure I-5. Active relay with storage of signals.

operational frequencies, since it does not require fulfillment of the condition of simultaneous visibility of the satellite from both of the corresponding points; the satellite altitude flight is not high.

Passive Relay of Signals by AES. In this case there is no radiotechnical equipment on board the AES. The radio waves radiated by the ground stations are reflected by the surface of the AES, which becomes the source of secondary radiation received by the ground receiving antenna.

It is obvious that the power requirements for radio links using passive relay through the AES is considerably higher than for radio links with active relaying. However, communication systems with passive relay do have a number of advantages, including (ref. 125):

(a) simplicity and reliability, associated with the absence of equipment o board the satellite (in the simplest case the AES is an inflatable metal sphere);

(b) flexibility with respect to the choice of the operating frequency and operating mode, requiring in case of necessity only an alteration of the operating conditions of the ground equipment;

(c) possibility of simultaneous relay of signals by various ground stations at different frequencies without creation of mutual interference.

4. Selection of Optimal AES Orbit for Creation of Global and Local Ground Communication Systems

The selection of the specific form of satellite orbit for the design of a ground communication system is determined by the problems which the communication system is called on to resolve. The entire spectrum of possible problems is divided into two major groups: the problem of the provision of worldwide communication (global communication systems) and the problem of communi- /15 cation within individual regions of Earth's surface (local communication systems).

Various methods of resolving both of these groups of problems are described in the literature. We shall consider some of them.

Global Communication System Based on Three Stationary Satellites. As we mentioned, the stationary satellite is the name given to the satellite in an equatorial orbit with a period of revolution equal to the period of rotation of Earth about its axis (T_c = 24 hr), and an altitude h = 36,000 km above the surface of Earth. For a ground observer this satellite will appear motionless. Three stationary satellites having equidistant spacing with intervals of 120° provide for "illumination" of 98 percent of the area of Earth's surface, with the exception of small regions around the poles (ref. 13). In order to cover the terrestrial hemisphere, the satellite antenna must have a beam width $\alpha = 18^\circ$. The maximal slant range from the ground station to the satellite is equal to 42,000 km.

In fig. I-6 the symbols C_1, C_2, C_3 denote the stationary satellites and P_1, P_2, P_3 denote the ground relay stations located on the boundaries of the individual zones of visibility for their communication with one another. In reference 13 it is indicated that with construction of relay ground stations in Africa, Tokyo and Houston and equidistant positioning of three satellites it is possible to accomplish long-range radio communication between practically all the countries of the world; using two satellites, one over the Atlantic Ocean and the other over the Pacific Ocean, it is possible to create a communication system which encompasses the majority of the capitalistic countries.

The creation of a communication system using stationary satellites will apparently encounter several difficulties of a practical nature (refs. 3 and 14),

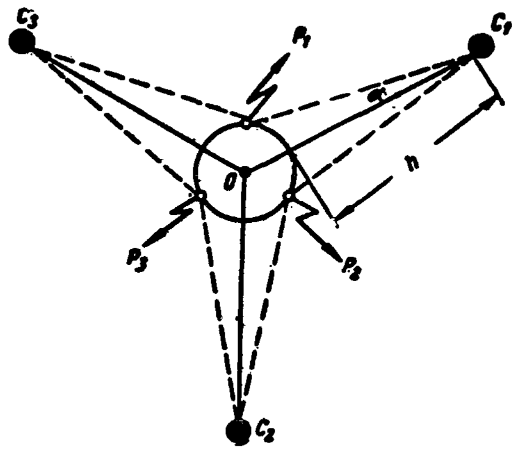


Figure I-6. System of three stationary satellites.

associated with the launch of a heavy satellite into high orbit and maintenance of its position in orbit (the gravitational disturbing action of the gravitational fields of the Moon and Sun will distort the satellite orbit) and with the necessity for precise orientation of the satellite antenna system toward Earth. Thus, for example, in reference 11 an accuracy of stabilization of the antenna system on board the satellite equal to ± 4° is assumed. /16

The U.S. specialists consider (ref. 3) that the state of rocket technology limits the magnitude of the useful load of the satellite which can be launched at the present time into a stationary orbit to a weight of the order of 180 kg. The distribution of the weight of such a satellite between the individual elements of system equipment designed for 96 telephone channels is shown in table I-1.

An idea of the possibilities of such a satellite is given by the data characterizing the variation of the average relay transmitter power as a function of the satellite weight (ref. 3). This relation is shown in figure I-7. /17
From the figure we see that with a repeater weight of about 30 kg the average output power will amount to about 3 W. In connection with the tendencies toward increase of the booster power, the weight of the repeater will go up. There will be a corresponding increase in the average output power of the transmitter.

We might also consider the variant of the global communication system with passive relay. In this case an inflatable metallized sphere of large diameter launched into a stationary orbit serves as the repeater. Among the drawbacks of this system is the presence of large disturbances of the orbit as the result of the pressure of the solar rays.

As an example of the solution of the problem of providing local communication, we shall present one of the possible variants of provision of communication for the portion of the northern hemisphere adjacent to the North Pole, and also the transatlantic communication system.

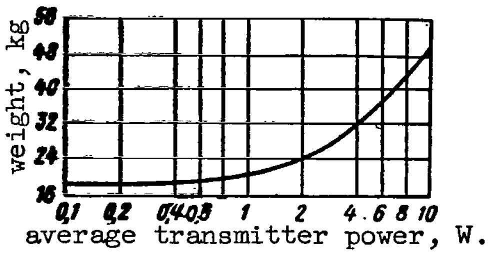


Figure I-7. Variation of transmitter power with repeater weight.

TABLE I-1

Equipment item	Item weight in % of total weight of satellite
Communications equipment	12
Stabilization system, including fuel supply	23
Power supply	39
Structure	26

Communication System for the Northern Hemisphere Based on Satellites in Inclined Elliptic Orbits. Reference 15 presents data on the geographical distribution of the population and telephones on the surface of Earth, excluding the USSR and China (fig. I-8).

We see from the figure that the communication requirements in the northern hemisphere are considerably higher than in the southern. We can also note that the provision of communication for the regions of Earth between 30 and 60°N resolves, basically, the problem of global communication as well. The com- /18 munication system proposed by the authors of reference 15 based on satellites in elliptic orbits with apogee in the northern hemisphere solves, in addition to the problem of providing communication between 30 and 60°N, also the problem of the communication in the northern polar basin, where the presence of zones of anomalous absorption does not permit reliable communication using shortwaves. From the point of view of the optimal radio illumination, it would be advisable to construct communication systems in polar orbits. But, as a result of the deviation of the gravitational field of Earth from centrality because of the oblateness of Earth, the apogee will drift in the orbital plane and after some time will be displaced into the southern hemisphere. Analysis of the movement of a satellite with account for the deviation of the gravitational field of /19 Earth from centrality, presented in reference 16, shows that the rate of drift of the apogee is equal to zero for an inclination of the orbit to the equatorial plane of 63°.

Calculations of the time of simultaneous visibility of the satellite from New York and London have shown that one satellite in an elliptical orbit inclined at an angle of 63°, with a period of revolution of 6 hr, apogee height of

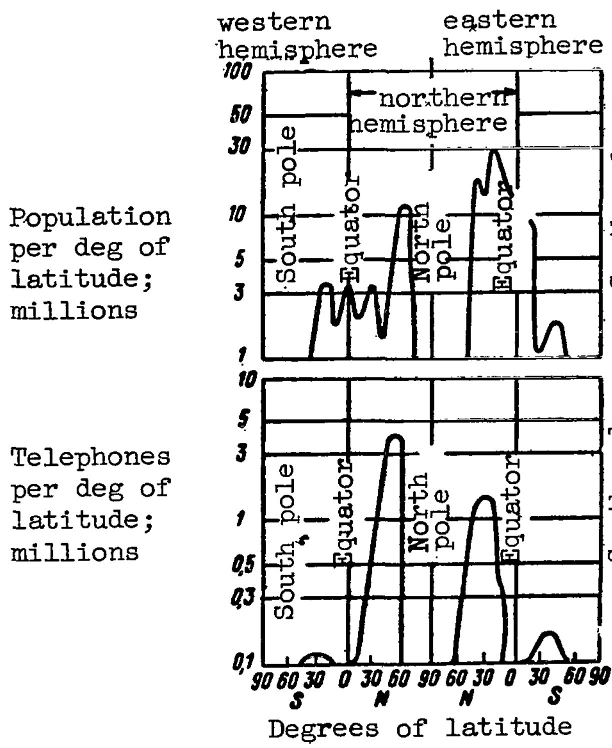


Figure I-8. World distribution of population and telephones.

12,600 miles (20,000 km), and a perigee height of 300 miles (500 km), is visible simultaneously at both points on the average about 9 hr a day (ref. 15). Four satellites, suitably phased, provide for almost around-the-clock radio illumination of all points of Earth's surface above 40°N. In addition, there is the possibility of maintaining intermittent (scheduled) communication via satellite even in regions of southern Asia.

In order to cover the terrestrial hemisphere, the satellite antenna must have a beam width, when the satellite is at the apogee, of $\alpha = 28^\circ$. The maximal slant range is 25,600 km.

Transatlantic Communication System Using Satellites in Circular Polar Orbits. Pierce and Kompfner (ref. 17) have analyzed the possibilities of creating a transatlantic communication system (the Newfoundland-Hebrides Islands line) with relay via AES. In order to determine the orbits, inclined or polar, which are optimal for this case, they carried out calculations of the time of simultaneous visibility of the satellite from two points: Newfoundland (48°N, 55°W) and Lewis Island (Hebrides, 50°N, 7°W) for satellite flight altitudes of 800, 1,600, 2,400, 3,200, 4,000 and 4,800 km.

Figure I-9 shows the zones of simultaneous visibility for the flight altitudes listed above, from which we can calculate the time of simultaneous visibility of the satellite from given points for various orbital altitudes as a function of the longitude of the orbits.

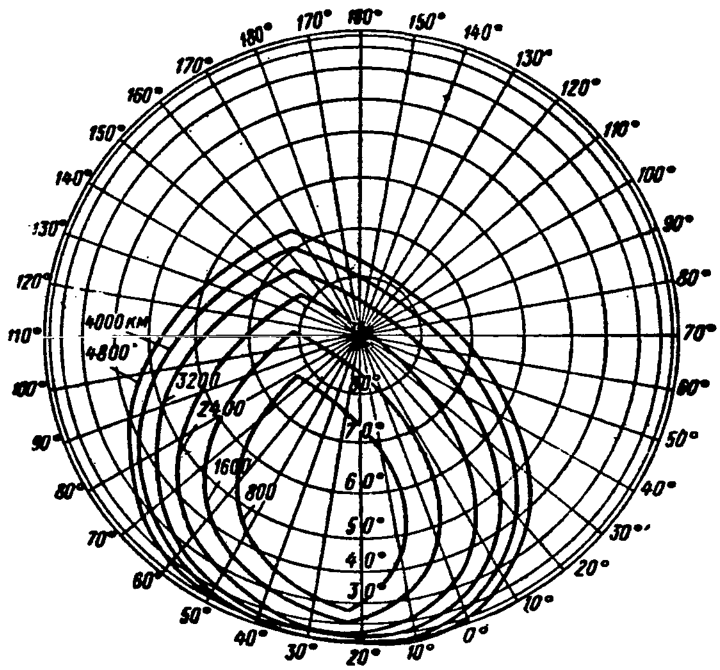


Figure I-9. Zones of simultaneous visibility of satellites in polar orbits at differing altitudes.

TABLE I-2

Parameter	Height above Earth's surface, km					
	800	1,600	2,400	3,200	4,000	4,800
Period of revolution T_c, min	100.4	118.0	136.0	155.0	175.2	195.2
Average time p, %	3.5	6.9	12.9	17.7	19.6	22.0

Computations were made for 72 polar orbits at intervals of 5° of longitude, and the average time of simultaneous visibility was determined in percent of the orbital period. The results of the calculations are presented in table I-2. (Optical visibility and elevation angle of the ground antenna of $\delta = 0^{\circ}$ were assumed.) /20

Similar calculations made for a satellite in an orbit inclined at an angle of 55° and an altitude of 3,200 km showed that the average time of visibility in this case is reduced from 17.7 to 10.2 percent. Thus, polar orbits are optimal for this case. /21

It was assumed that the elevation angle of the ground antenna was $\delta = 0^{\circ}$. However, as a result of Earth and atmospheric interferences, particularly important in the case of a sensitive receiver with parametric or molecular amplifiers, and also as a result of anomalously high refraction and the danger of fading because of multipath reception due to the reflection from the surface of Earth, the antenna must be raised somewhat above the horizon. In this case the time of simultaneous radio visibility of the satellite will be reduced. Calculations were made of the simultaneous radio visibility of a satellite with a flight altitude of 4,800 km for various antenna elevation angles: $\delta = 0$, 3.25, 7.25, 12.6°. (The magnitudes of δ were selected from the condition that with these values of the elevation angle of the antenna the radio visibility of a satellite at an altitude of 4,800 km will be the same as the optical visibility of a satellite with altitudes of 4,000, 3,200, 2,400 km, respectively.) The results of the calculations are presented in table I-3.

We see that with an antenna elevation angle $\delta = 7.25°$ the radio visibility of the satellite is reduced from 22 percent with $\delta = 0°$ to 17.7 percent.

For 24-hour or nearly 24-hour communication a definite number of satellites is required. Here we can consider two cases:

(1) The satellite orbits are synchronous relative to one another.

With suitable relative positioning of the satellites, practically around-the-clock communication will be provided, if we include in the system $n = 100\%/p$ satellites, where p is the average time of radio visibility in percent. Table I-4 presents the data on the required number of phased satellites for several altitudes with $\delta = 0$, and table I-5 the same for an altitude of 4,800 km with various δ. /22

(2) The satellite orbits are identical but not synchronous, and the satellites have random distribution in space and time.

If f is the portion of the period during which the satellite is simultaneously visible by both correspondents (averaged over a large number of orbits), then (1-f) is the probability that the satellite will not be seen by the correspondents. For n satellites the probability that none of them will be seen is equal to

$$i = (1 - f)^n. \tag{I-2}$$

Whence, assuming a definite value of the communication reliability 1-i, we find the number of satellites required to provide this reliability

TABLE I-3

Antenna elevation angle δ, deg	0	3.25	7.25	12.6
Average time p, %	22	19.6	17.7	12.9

$$n = \frac{\log i}{\log (1-f)}. \qquad (I-3)$$

TABLE I-4

Flight altitude h, km	800	1,600	2,400	3,200	4,000	4,800
Number of satellites n	28.6	14.5	7.75	5.65	5.1	4.5

TABLE I-5

Antenna elevation angle δ, deg	0	3.25	7.25	12.6
Number of satellites n	4.5	5.1	5.65	7.75

The authors of reference 17 made the calculations for the number of satellites n with different values of 1-i for different flight altitudes with $\delta = 0$ and for the single height h = 4,800 km with differing δ. The results of the calculation are summarized in tables I-6 and I-7.

Comparison of the data of tables I-4, I-5, I-6 and I-7 shows that synchronization of the orbits gives a very considerable gain in the number of satellites required for the organization of reliable communication. /23

The further analysis included only the h = 4,800 km altitude. The field of view of Earth's surface from a satellite at this flight altitude is equal to $\alpha = 70^{\circ}$. The maximal slant range $R_{max} = 9{,}200$ km.

All these communication systems belong to the class of systems using immediate relay, for which a necessary condition is that of simultaneous visibility of the satellite from the corresponding points between which the communication is to be established. Several authors have proposed systems with a delay of the signal, in which the information is transmitted from the ground station at the moment of passage of the satellite over it, is recorded and then is reproduced at the time of passage of the satellite over the intended receiving station (refs. 18 and 19). One of the possible variants which has been suggested (ref. 14) is the active relay system with delay of the signal which is based on satellites launched into equatorial circular orbit with an altitude of about 2,000 km. The field of view of Earth's surface from the satellite /24 is equal to $\alpha = 100^{\circ}$; the maximal slant range is equal to $R_{max} = 5{,}400$ km.

Based on this consideration of the various methods of constructing ground communication systems using AES, we can formulate certain general principles which must be followed in the selection of the communication satellite orbits.

In the general case the problem is posed thusly: we are required to provide 24-hour communication over a certain area of Earth's surface.

TABLE I-6

Altitude, km	Number of satellites n			
	1-i = 10%	1-i = 1%	1-i = 0.1%	1-i = 0.01%
800	62.5	125	187	250
1,600	32.3	64.6	96.9	129.2
2,400	16.7	33.3	50	66.6
3,200	11.8	23.6	40	47.2
4,000	10.5	21	31.3	42
4,800	9.0	18.5	27.5	37

TABLE I-7

δ, deg	Number of satellites n		
	1-i = 10%	1-i = 5%	1-i = 1%
0	9	12	19
3.25	11	14	21
7.25	12	15	24
12.6	17	22	33

It is evident that the choice of the optimal orbit depends on the area of the territory, the geographical location (high, medium or low latitudes) and the configuration (the extent in longitudinal or latitudinal directions). It was shown that the communication systems using satellites in synchronized orbits is considerably more economical than the communication systems using satellites in random orbits, since in the first case the number of satellites required for communication is considerably less than in the second, particularly if we are required to provide a communication reliability close to 100 percent. Therefore, in the future we shall consider only the case of the synchronized orbits.

With the use of the synchronized orbits we can suggest the following simple method of design of a satellite system providing 24-hour radio illumination of a given territory. Let us calculate the time of simultaneous radio visibility of a single satellite from the given territory during a large number of orbits. We shall find the minimal time of radio visibility of the satellite Δt_{min} (in

the ith orbit) and determine the satellite coordinates for this orbit relative to Earth's surface, and also the time of radio visibility and radiorise and radioset of the satellite in the given territory. It is evident that 24-hour radio illumination of the given territory will be provided by a system of n satellites, where

$$n = \frac{T_c}{\Delta t_{min}} ,$$

where T_c is the period of revolution of each of the satellites following one after the other at equal intervals of time Δt_{min} and phased so that in the worst orbit (minimal time of radio visibility) at the moment of radioset of the kth satellite of the system the (k + 1)th satellite is rising. The necessary condition for the creation of such a satellite system is a nonzero value of the minimal time of simultaneous radio visibility Δt_{min} of one satellite of the system from the given territory. For given configuration and size of the "illuminated" territory this condition imposes a limitation on the height of the orbit--the orbital altitude must be greater than some minimal value for which $\Delta t_{min} = 0$. /25

With increase of the orbital height, the zone of radio illumination by the satellite of the territory of Earth's surface expands, and the time of minimal simultaneous radio visibility of the satellite from the given territory Δt_{min} increases. According to Kepler's third law the period of the revolution T_c also increases. The time of minimal radio visibility Δt_{min} depends not only on the orbital height, but also on the shape of the orbit (magnitude of its eccentricity), and its orientation relative to Earth (polar, inclined, equatorial). Thus, the number of satellites in the system n is a complex function, depending on several variables.

From the point of view of reduction of the economic expenditures on the creation of the satellite system and convenience of handling, it is desirable to have the minimal possible number n_{min} of satellites in the system. Under this condition, the costs of satellite launches are reduced, the time of simultaneous radio visibility of a single satellite is increased, which alleviates the problem of tracking the satellite by the ground antennas and reduces the number of interruptions in communication (with the "swing" of the antenna from the setting satellite to the "rising"); the problem of synchronizing the satellite orbits is also eased. From the point of view of reducing the power required for the radio link it is desirable to reduce the orbital height, which can be done by increasing the number of satellites. Consequently, the solution can only be a compromise and must take account of the various aspects (economic, power consumption, equipment and others) of the creation of the entire communication system as a whole, and not simply the system of satellites.

We will show that with a high degree of accuracy of orientation of the satellite antenna pattern toward the center of Earth the required satellite transmitter radiation power increases only slightly with an increase of the flight altitude. The increase of range is compensated by a corresponding increase of the amplification factor of the transmitting antenna, since with increase of the altitude there is a reduction of the field of view of Earth's surface from the satellite. Consequently, with a simple and reliable stabilization system on board the satellite, which provides a high degree of accuracy of /26 orientation of the satellite in space with small energy consumption (for example, a system of spatial orientation using the gradient of Earth's gravitational field), the optimal orbit will be that for which the number of satellites in the system is minimal ($n = n_{min}$).

For 24-hour radio illumination of large territories adjacent to the equator it is advisable to use stationary satellites. One stationary satellite located over the Atlantic Ocean will provide 24-hour radio illumination of the territory of all countries of South America, Western Europe, the countries of Africa (with the exception of Somali and part of Abyssinia), the territory of the USA (with the exception of Alaska) and part of the territory of Canada and Greenland. One stationary satellite located over the Indian Ocean will provide 24-hour radio illumination of the major portion of the territory of the USSR (with the exception of the islands in the North Arctic Ocean and Chukotka), the territory of the countries of Eastern Europe, Asia Minor, India, China, Australia and part of Africa.

A major advantage of the stationary satellites is their "immobility" relative to the ground observer, which eliminates the problem of tracking the satellite. In addition, fixed antennas can be made very efficient. Their disadvantages were noted before (small payload in orbit, high demands on the orientation accuracy, disturbances of the orbit as the result of the action of the gravitational fields of the Moon and Sun).

In certain cases the need arises for provision of 24-hour communications for the territories lying near Earth's poles. We have shown the method of designing a global communication system based on three equidistantly spaced stationary satellites. A disadvantage of this communication system is that the regions adjacent to Earth's poles are not illuminated. The upper limit of the zone /27
of radio visibility of the system of three satellites is 76.5°N with $\delta_{min} = 5°$; the lower limit of the zone is the latitude of intersection of two neighboring zones of radio visibility of individual satellites, equal to 62°N with $\delta_{min} = 5°$. Thus, a portion of the mainland territory of the USSR, the islands in the North Arctic Ocean and all of the North Polar Basin are eliminated from the overall communication system.

In connection with the increase of international traffic, there is increased interest in the North Polar Basin, since the air route across the North Pole is the shortest distance from the USSR to North America (Canada, USA). We mentioned previously that the shortwave radio circuits, passing through the zone of polar absorption or the links lying within this zone, are frequently subjected to communication breakdown and are not suitable for provision of communications with a reliability of 100 percent. This problem can be resolved most simply by satellites.

It is easy to show that for the design of a system of satellites providing 24-hour radio illumination of the territory to the north of some latitude φ_0 (in the present case $\varphi_0 = 62°N$) it is advisable to use satellites in elliptic orbits with an apogee in the northern hemisphere. According to Kepler's second

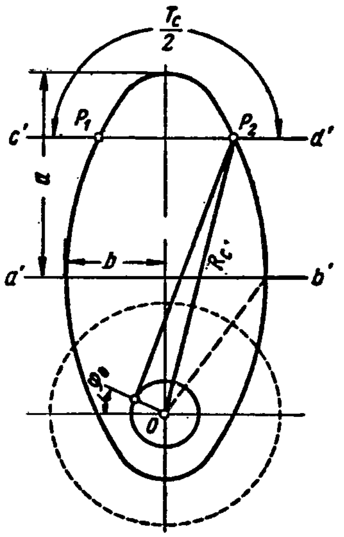


Figure I-10. System of two satellites in elliptic orbits.

law (law of areas) a satellite travels considerably faster in the region of the perigee than in the region of the apogee. Consequently, the satellite is longer in the apogee, and the line of time symmetry c'd' is shifted relative to the line of symmetry of the ellipse a'b' (fig. I-10).

It is evident that if we launch two satellites into an elliptic orbit /28 and phase them so that at the moment of setting of the first satellite at the point P_2 of the orbit the second satellite rises at the point P_1 of the orbit (satellites separated by a time interval equal to half of the period T_c), we can provide 24-hour radio illumination on the territory to the north of some latitude φ_0. The advantage of this communication system is the absence of ground relay stations. With the use of circular orbits the analogous problem is solved by the presence in the system of a minimum of three satellites, because in this case the line of time symmetry coincides with the line of symmetry of the orbit, and the time of radio visibility of the satellite is always less than half the period of revolution T_c.

We should note that the construction of a communication system to the north of some latitude φ_0 using satellites in elliptic orbits is advantageous not only from the economic point of view, because the number of satellites in the system is minimal, but also from the energy point of view, since the maximal height of the satellite orbit (apogee height) in the elliptic orbit in the system of two satellites is less than the height of the orbit (distance to the surface of Earth) of the satellite in a circular orbit in the system with three satellites. Calculations show that the problem of providing 24-hour radio coverage of the territory to the north of $\varphi_0 = 62^\circ$ is solved by two satellites in an elliptic orbit with inclination $i = 64^\circ$, perigee height $h_p = 500$ km and apogee height $h_a = 23{,}000$ km (the value $i = 64^\circ$, close to $i = 63.5^\circ$, was chosen to prevent drift of the apogee in the orbital plane due to the deviation of the gravitational field of Earth from centrality as the result of its oblateness; with $i = 64^\circ$ the displacement of the apogee after one year will not exceed 5°) (refs. 13, 15 and 125). The minimal elevation angle of the ground antenna δ_{min} was taken as 10°. With this value of elevation angle δ_{min} the system of three satellites in circular orbit cannot solve the stated problem with any altitude of flight. This problem is solved by a system of three satellites in a circular polar orbit, if we set $\delta_{min} = 0$ and launch the satellites into an orbit with an altitude $h = 190{,}000$ km. The problem of radio coverage of the given territory with $\delta_{min} = 10^\circ$ is solved by a system of four satellites in circular polar orbit with $h = 45{,}000$ km. This comparison gives a vivid illustration of the advantages /29 of the elliptic orbit in comparison with the circular.

We can draw the following conclusions

(1) The selection of the optimal orbit of the satellite must be performed separately for each specific communication system and must take account of such factors as the geographic location and dimensions of the territory for which the given system is intended; the cost of the satellite system and the cost of the ground equipment; the capabilities of the onboard power sources; the reliability of the onboard equipment, payload in orbit, etc. (ref. 81).

(2) for provision of 24-hour communications over large territories near the equator it is advisable to make use of stationary satellites (ref. 13);

(3) for provision of 24-hour communication in the regions near the North Pole it is advisable to use satellites in elongated elliptical orbits with apogee in the northern hemisphere (ref. 15).

In this discussion we have used the concept of the time of simultaneous radio visibility of the satellite from the ground. We have shown that this parameter is the determining factor in the selection of the optimal orbit. In the following pages we shall consider the method of calculating the time of simultaneous radio visibility of the satellite from some territory of Earth's surface.

A satellite is "visible" simultaneously from some territory if it is "visible" from any point of this territory. For the calculation of the time of simultaneous visibility of a satellite from some territory it is necessary to determine for each orbit the moments of radiorise and radioset of the satellite for all the points of the given territory, and the difference of the times of the latest radiorise and the earliest radioset will be the time of simultaneous radio visibility of the satellite from the entire territory.

It is obvious that it makes sense to carry out these calculations only for the points on the boundary of the territory. Depending on the configuration of the territory, in practical calculations we usually limit ourselves to 4, 3 or even 2 points. We shall derive the equations for the determination of the time of radio visibility of a satellite from some point of observation S (fig. I-11).

The satellite is "visible" (radio visibility) from an observation point, if the condition $\angle PSO \geq 90^{\circ} + \delta_{min}$ is satisfied, where δ_{min} is the minimal elevation angle of the transceiving station antenna located at point S. It /30 is evident that in the case of equality, when $\angle PSO = 90^{\circ} + \delta_{min}$, the distance between the satellite and the observation point is maximal.

We obtain the equation for the maximal slant range from the triangle PSO by using the theorem of cosines

$$R^2_{max}(t) + 2R_{max}(t)\, R_0 \sin \delta_{min} + R_0^2 - R_c^2(t) = 0, \qquad \text{(I-4)}$$

where $R_{max}(t)$ is the maximal distance between the satellite and the observation point (varies in the course of time t);
$R_c(t)$ is the radius vector of the satellite, measured from the center of Earth (varies with time t);
R_0 is the radius of Earth.

Solving equation (I-4) relative to $R_{max}(t)$, we find

$$R_{max}(t) = \sqrt{R_c^2(t) - R_0^2 + R_0^2 \sin^2 \delta_{min}} - R_0 \sin \delta_{min} \qquad \text{(I-5)}$$

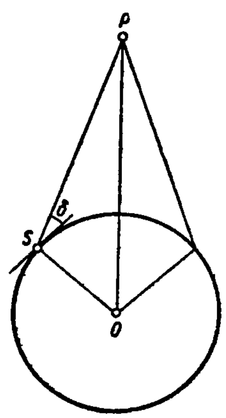


Figure I-11. Geometry of problem on determination of satellite equivisibility.

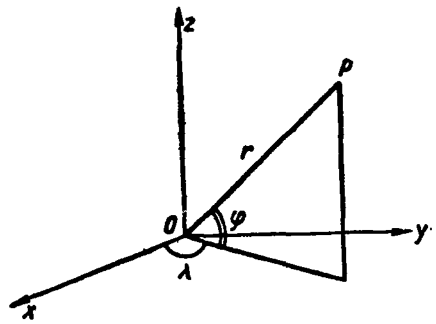


Figure I-12. Coordinate system for determination of satellite equivisibility.

Thus, at the instant of time t the satellite is visible from the observation point S, if the condition $R(t) \leq R_{max}(t)$ is satisfied. Let us find the variation of R(t) with time. It is known that in the Cartesian coordinate system x, y, z the distance between points with coordinates x_2, y_2, z_2 and x_1, y_1, z_1 is equal to

$$R=\sqrt{(x_2-x_1)^2+(y_2-y_1)^2+(z_2-z_1)^2}. \tag{I-6}$$

In the spherical coordinate system r, φ, λ (fig. I-12) fixed with the center of Earth (latitude φ is measured from the equatorial plane, longitude λ from the Greenwich meridian) the Cartesian coordinates x, y, z are written in the form /31

$$x=r\cos\varphi\cos\lambda,$$
$$y=r\cos\varphi\sin\lambda,$$
$$z=r\sin\varphi.$$

Substituting in (I-6) in place of x_2, y_2, z_2, x_1, y_1, z_1 the expressions R_c, φ_c, λ_c, R_0, φ_0, λ_0, we obtain

$$R(t)=$$
$$=\sqrt{R_c^2(t)+R_0^2-2R_0R_c(t)\cos\varphi_c(t)\cos[\lambda_c(t)-\lambda_0]\cos\varphi_0-2R_c(t)\sin\varphi_c(t)\sin\varphi_0R_0}. \tag{I-7}$$

We see that in order to clarify the variation with time of the functions R(t) and $R_{max}(t)$ we must know the variation with time of the coordinates of the satellite $R_c(t)$, $\varphi_c(t)$, $\lambda_c(t)$.

The most general case is that of the motion along an inclined elliptic orbit. Let us consider this case. It is known that the location of a satellite in the orbit is determined by the radius vector R_c and the value of the instantaneous anomaly f--the angle between R_c and the axis of the apsides (fig. I-10).

The variation of the true anomaly f with time is determined by means of the solution of the transcendental equation (ref. 20)

$$\frac{2\pi}{T_c}t = \arccos\left[\frac{1}{e} - \frac{1-e^2}{e(1+e\cos f)}\right] - e\sqrt{1-\left[\frac{1}{e} - \frac{1-e^2}{e(1+e\cos f)}\right]^2}, \tag{I-8}$$

where e is the eccentricity of the ellipse $\left(e=\frac{\sqrt{a^2-b^2}}{a}\right)$ with semimajor and semiminor axes equal to a and b, respectively;

$2\pi/T_c$ is the average angular velocity of the satellite (ω_c) in the orbit with the period of revolution T_c. Time t is reckoned from the time of passage of the satellite through the perigee.

From the known value of the true anomaly f we find the corresponding value of the radius vector R_c /32

$$R_c = \frac{a(1-e^2)}{1+e\cos f}. \tag{I-9}$$

Assuming that the angle of inclination of the orbit to the equatorial plane of Earth is equal to i and that the longitude of the perigee at the moment t = 0 is equal to λ_{c0}, from geometric considerations we obtain the following equation for the instantaneous latitude φ_c and longitude λ_c of the satellite

$$\varphi_c(t) = \arcsin[-\cos f \sin i], \tag{I-10}$$

$$\lambda_c(t) = \lambda_{c0} - \omega_e t + \pi - \mathrm{arctg}[-\mathrm{tg}\, f \sec i], \tag{I-11}$$

where ω_e is the angular velocity of rotation of Earth about its axis, equal to 0.25 deg/min.

Example. With $0 < f < 2\pi$

$$180° > \mathrm{arc\,tg}[-\mathrm{tg}\, f \cdot \sec i] > -180°.$$

If the orbit is circular, then $f = \omega_c t$, $R_c = R_0 + h$, where h is the height of the orbit above the surface of Earth, and consequently

$$\varphi_c(t) = \arcsin[-\cos\omega_c t \sin i], \tag{I-10'}$$

$$\lambda_c(t) = \lambda_{c_0} - \omega_з t + \pi - \operatorname{arc\,tg}[-\operatorname{tg}\omega_c t \sec i] \tag{I-11'}$$

The equations for φ_c and λ_c (and, consequently, also for R and R_{max}) have the simplest form in the case of polar and equatorial circular orbits.

Polar orbits($i = \pi/2$)

$$\left.\begin{aligned}\varphi_c &= -\frac{\pi}{2} - \omega_c t,\\ \lambda_c &= \lambda_0 - \omega_з t\end{aligned}\right\}$$

at the moment t = 0, the satellite is over the South Pole.

Equatorial orbits ($i = 0$)

$$\left.\begin{aligned}\varphi_c &= 0,\\ \lambda_c &= \lambda_{c_0} - \omega_з t - \omega_c t.\end{aligned}\right\}$$

Analysis of equations (I-8), (I-9), (I-10) and (I-11) shows that for the determination of the time variation of the satellite coordinates R_c, φ_c and λ_c, it is necessary to know the following orbital parameters: value of the semi- /33 major axis a, eccentricity e, angle of inclination of the orbital plane to the equatorial plane i.

It is evident that the semimajor axis a and the eccentricity e are known, if the altitudes of the apogee h_a and perigee h_p are known

$$a = R_0 + \frac{h_a + h_p}{2}, \quad e = \frac{h_a - h_p}{2\left(R_0 + \frac{h_a + h_p}{2}\right)}.$$

To find the times of radiorise and radioset of the satellite at the point of observation with the coordinates R_0, φ_0 and λ_0, we construct on a single drawing the graphs of the relations $R(t)$ and $R_{max}(t)$, and from the condition $R(t) \leq R_{max}(t)$ we find the time of radio visibility of the satellite from the observation point S after the ith orbit.

We proceed similarly to calculate the time of simultaneous radio visibility of the satellite from two or more observation points. For each observation point we calculate the variation with time of the distance from the point to the satellite, and on one drawing we construct the plots of $R_1(t)$, $R_2(t)$, ..., $R_k(t)$ and $R_{max}(t)$.

From the condition that simultaneously

$$\begin{gathered} R_1(t) \le R_{max}(t), \\ R_2(t) \le R_{max}(t), \\ \cdots\cdots\cdots \\ \cdots\cdots\cdots \\ \cdots\cdots\cdots \\ R_k(t) \le R_{max}(t), \end{gathered}$$

we find the time of simultaneous radio visibility of the satellite Δt from the observation points S_1, S_2, ..., S_k.

5. Power Analysis of Radio Link with Retransmission Via AES

Since the power capability of the transmitter on Earth is considerably greater than that of the transmitter on board the satellite, we shall consider only the satellite-Earth radio link.

In the case of active retransmission, the communication equation is written in the form (refs. 121 and 125) /34

$$P_t = N_0 k T_{\text{eff}} F \frac{16R^2}{G_1 \eta D_2^2} \Gamma N_1; \qquad \text{(I-12)}$$

in the case of passive retransmission the communication equation (radar transmission equation) has the form

$$P'_t = N_0 k T_{\text{eff}} F \frac{R_1^2 R_2^2 64\lambda^2}{S_0 D_{21}^2 D_{22}^2} \Gamma_1 \Gamma_2 N_1, \qquad \text{(I-13)}$$

where P_t is the required satellite transmitter radiation power;

P'_t is the required ground transmitter radiation power;

N_0 is the receiver RF signal/noise ratio (ahead of the detector);

k is the Boltzmann constant;

T_{eff} is the effective noise temperature at the receiver input;

F is the RF passband of the ground receiver (ahead of the detector);

G_1 is the satellite transmitter antenna gain;

D_2 is the diameter of the ground receiving antenna;

D_{21} is the diameter of the ground transmitting antenna with passive retransmission;

D_{22} is the diameter of the ground receiving antenna with passive retransmission;

R is the slant range from the satellite to the ground correspondent with active relay;

R_1 is the slant range from the satellite to the transmitting station on the ground;

R_2 is the slant range from the satellite to the receiving station on ground;

Γ is a safety factor which accounts for the absorption and fading of the signal in the ionosphere and the absorption in troposphere;

Γ_1, Γ_2 same, but for the case of passive relaying (Γ_1--on the path from the ground station to the satellite; Γ_2--on the path from the satellite to the ground receiving station);

S_0 effective scattering area of the satellite in the direction towards the receiving antenna with passive relaying; /35

η antenna area use factor ($\eta \approx 0.5$);

λ operating wavelength;

N_1 margin for loss due to nonuniformity of directivity pattern, for loss in antenna feeder system, polarization losses.

Let us make an analysis of the various quantities appearing in the communication equation.

It is clearly sensible to calculate the power requirements of the radio links, with both active and passive relay, for the worst case, when the slant range between the corresponding stations (ground station-satellite) is maximal. Consequently, we must take

$$R_1 = R_2 = R = R_{max}$$

We can show that with the satellite at an altitude h above Earth's surface the maximal slant range with an elevation angle of the ground antenna of δ_{min} is equal to

$$R_{max} = \sqrt{(R_0 + h)^2 - R_0^2 \cos^2 \delta_{min}} - R_0 \sin \delta_{min} \qquad \text{(I-14)}$$

where R_0 is the radius of Earth.

Note. For the case of the elliptic orbit the energy calculation must be performed for $h = h_a$, where h_a is the apogee altitude.

Since the gain G_1 of the transmitting antenna on board the satellite is uniquely related with the width of the antenna pattern, evidently it must be constant and must be determined by the field of view of Earth's surface from the satellite α with account for the correction for the instability of the orientation of the satellite towards the center of Earth γ. Consequently, the width of the antenna pattern to the half-power point must be equal to $\theta = \alpha + 2\gamma$, where α is the field of view of Earth's surface from the satellite, associated with the satellite flight altitude h by the following relation

$$\alpha = 2 \arcsin \left[\frac{R_0}{R_0 + h} \cos \delta_{min}\right]. \qquad \text{(I-15)}$$

The average gain of an antenna with width of directivity pattern to the half-power points θ is equal to /36

$$G = \frac{2}{1 - \cos\frac{\theta}{2}}. \qquad \text{(I-16)}$$

Substituting in (I-16) the expression for θ in terms of α and γ, we obtain the following final expression for G_1

$$G_1 = \frac{2}{1 - \cos\left[\arcsin\left(\frac{R_0}{R_0 + h} \cos \delta_{min}\right) + \gamma\right]}. \qquad \text{(I-17)}$$

With travel of the satellite in the limits of the visibility of its ground receiving station, the elevation angle of the ground antenna varies in the range $\delta_{min} < \delta < 90^{\circ}$. The slant range (distance between the satellite and the ground station) varies in the range $R_{max} > R > h$. Although the radiation power of the onboard transmitter is constant, the signal/noise ratio at the input of the ground receiver N_0 will still be variable. Here the differential between the maximal and minimal values of N_0 will be equal to R^2_{max}/h^2 (with constant gain of the transmitting antenna). Evidently, in order to maintain a constant signal/noise ratio at the input of the ground receiver it is necessary suitably to vary the power of the radiation of the onboard transmitter, or to have a transmitting antenna with a directivity pattern determined by the satellite flight altitude h and the accuracy of orientation of the satellite toward the center of Earth γ. In the case of a circular satellite orbit with $\gamma = 0^o$, the function describing the shape of the directivity pattern has the form $\Phi(\alpha) = R(\alpha)/R_{max}$.

As the result of the spatial redistribution of the power radiated by the transmitter, the signal/noise ratio N_0 for the case $R = R_{max}$ increases, while for the case $R = R_{min}$ it diminishes. Since the design of the radio link is made for the case $R = R_{max}$, the use of a transmitting antenna with a special shape of directivity pattern gives a power gain (voltage gain) equal, in the best case, to /37

$$n_{max} = \frac{1-\cos\frac{\theta}{2}}{\frac{R_0+h}{4R_{max}}(1-\cos\theta)-\int\limits_0^{\frac{\theta}{2}}\frac{R_0}{R_{max}}\sqrt{1-\left(\frac{R_0}{R_{max}}\right)^2\sin^2 x}\,\sin x\,dx} \quad \text{(I-18)}$$

It should be noted that in practice the energy gain n will be less than n_{max} because of the impossibility of constructing an antenna with a directivity pattern exactly described by the equation for $\Phi(\alpha)$. Since the exact value of the gain n is indeterminate, we shall not take account of it in the energy calculations, and the realization of such a gain will give some margin in the power of the radio link for the case of active relaying. In this case a marked power effect is obtained only in the case of low-flying satellites, with $h = 2{,}000$ km, $n \approx 2$.

The value of the RF signal/noise ratio N_0 appearing in the communication equation (I-12) is determined by the required quality of the transmisssion and the form of modulation or keying used.

For space radio links, characterized by long communication distances and limited power potential of the onboard repeater, the question of the choice of the optimal form of modulation becomes particularly acute. For this reason we shall study it in more detail.

Comparison of Various Methods of Information Transmission. There are various criteria for the comparison of the methods of transmitting information. We shall make use of the criteria used in reference 21. Here two characteristics are taken as primary for the communication system--the noise immunity and the efficiency.

Noise rejection of a communication system is the term given to the capability of the system to oppose the harmful effect of interference on the transmission of information. The quantitative measure of the noise immunity is the degree of correspondence of the received information to that transmitted, /38 i.e., the accuracy of the reproduction of the information at the point of reception.

With the transmission of continuous information, the criterion of the accuracy of the reproduction can be the mean square error

$$\varepsilon^2 = \overline{[u(t) - v(t)]^2}, \tag{I-19}$$

where u(t) and v(t) are, respectively, the ensemble of the transmitted information and of the received information at the point of reception.

The relative error is defined as the ratio

$$E = \frac{\varepsilon^2}{P_m} = \frac{\int\limits_0^{F_m} \sigma_m^2(f)\, df}{\int\limits_0^{F_m} P_m(f)\, df} = \frac{S_m}{P_m}, \tag{I-20}$$

where $\sigma_m(f)$ is the noise intensity at the output of the receiver at the frequency f;

F_m is the receiver low-frequency passband;

P_m is the average signal power at the receiver output.

Thus, the relative error is completely defined by the ratio of the average noise power S_m at the output to the average information power P_m.

For practical calculations the criterion of noise immunity can be taken to be the gain B and the generalized gain B' of the system

$$B' = \frac{N'_m}{N'_0}, \quad B = \frac{N_m}{N_0}, \tag{I-21}$$

where

$$N'_m = \frac{P_m F_m}{S_m}; \quad N_m = \frac{P_m}{S_m};$$

$$N_0 = \frac{P}{S}; \quad N'_0 = \frac{PF}{S};$$

S is the noise power at the receiver input;
F is the frequency band in which the noise power at the receiver input is measured;
P is the signal power at the receiver input;
N_0, N_m are the signal/noise ratios at the input and output of the receiver, respectively. /39

The quantity B' completely characterizes the noise immunity, since, with a given spectral density of the noise at the input $\sigma^2 = S/F$ and a given power of the input signal P, it makes possible a unique determination of the signal/noise ratio at the output. The larger B', the less the power of the input signal P required to obtain the given quality of the output signal with the given noise intensity σ^2.

With the retransmission of discrete signals, the accuracy of the reproduction is determined by the error probability P_{err}.

For the evaluation of the efficiency of the communication system, reference 21 introduces a system of coefficients characterizing the utilization of the basic parameters of the channel. These coefficients are the signal power use factor--the β-efficiency,

$$\beta = \frac{V\sigma^2}{P} = \frac{V}{N_0 F} = \frac{VB}{FN_m} \tag{I-22}$$

and the channel frequency band use factor--the γ-efficiency,

$$\gamma = \frac{V}{F}, \tag{I-23}$$

where V is the rate of transmission of information in binary units per second.

In the case of space radio communication links, the power of the transmitter is severely limited and the β-coefficient is the most important characteristic for the analysis of the various methods of transmission of information.

The upper limit for β can be obtained from the Shannon equation (ref. 22) for the limiting handling capacity of the channel C with additive noise in the form of normal white noise

$$V \leqslant C = F \log\left(\frac{P}{\sigma^2 F} + 1\right). \tag{I-24}$$

Whence

$$\beta_0 = \frac{\gamma}{2\gamma - 1}, \tag{I-25}$$

where β_0 is the limiting value for an arbitrarily small error probability.

The upper limit of β_0 is equal to $1/\ln 2$ with $\gamma \to 0$ ($F \to \infty$). The value $1/\ln 2$ is the absolute upper limit for β and cannot be exceeded in any real system with additive white noise. /40

In those cases when the peak signal power is limited rather than the average signal power (for example, in cases of passive retransmission, pulse modulation methods), it is advisable to use the peak β-efficiency for the characteristic of the system efficiency

$$\beta_{peak} = \frac{V}{\frac{P_{peak}}{\sigma^2}}, \qquad \text{(I-22')}$$

P_{peak} is the signal peak power.

Reference 22 introduces still another parameter--the channel capacity use factor, or the η-efficiency

$$\eta = \frac{V}{C}. \qquad \text{(I-26)}$$

According to Shannon the channel capacity in the general case is determined by the relation

$$C = \max V = \max \frac{H(A) - H_x(A)}{T}, \qquad \text{(I-27)}$$

where $H(A)$ is the entropy of the signal source;

$H_x(A)$ is the equivalent noise entropy.

For the continuous channel with fluctuating noise with uniform spectrum the relation (I-24) is used, written in the form

$$C = F \log\left(\frac{P}{S} + 1\right). \qquad \text{(I-28)}$$

On the assumption that the transmitted information has the statistical structure of white noise, the transmission rate is determined by equation

$$V = F_m \log\left(\frac{P_m}{S_m} + 1\right), \qquad \text{(I-29)}$$

where F_m is the frequency band of the transmitted information.

The information transmission rate in a system with L discrete levels, in which all levels have identical transmission probability and identical error probability, is determined by expression /41

$$V=\frac{1}{T}[\log L-P_{err}\log(L-1)+P_{err}\log P_{err}+(1-P_{err})\log(1-P_{err})], \tag{I-30}$$

where P_{err} is the error probability;

T is the signal duration.

The expression for the η-efficiency of discrete communication systems can be written in the form

$$\eta=\eta_1\eta_2=\frac{V}{V_0}\cdot\frac{V_0}{C}, \tag{I-31}$$

where V_0 is the transmission rate with ideal coding.

The quantity η_1, equal to V/V_0, can be termed the efficiency of the coding system, and the quantity η_2, equal to V_0/C, is termed the efficiency of the modulation system.

The channel parameter use factors are interconnected with one another by the following relations

$$\gamma=\beta\frac{P}{S}, \tag{I-32}$$

$$\eta=\frac{\gamma}{\log\left(\frac{\gamma}{\beta}+1\right)}. \tag{I-33}$$

The efficiency and the noise immunity are the most important criteria of the operation of the communication system in the presence of noise. We shall consider the best system to be that one which has the highest efficiency for a given noise factor, or, on the other hand, the one having the highest noise factor for a given efficiency (ref. 23).

Using these criteria, we shall make an analysis of the various methods of information transmission (digital and continuous) from the point of view of their suitability for space radio communication links.

For the transmission of discrete information, the most widely used types of keying are frequency (FT), phase (PT) and pulse-time (PTT). Also /42 promising are the multiple-stable-state systems with the use of orthogonal signals.

In the case of the reception of binary signals with amplitudes A_1, A_2, duration τ_0, having identical energy, with duration of the observation interval T, the error probability of the ideal receiver is determined by equation

$$P_{err} = \frac{1}{2}[1 - \Phi(h_0\sqrt{1-\rho})], \tag{I-34}$$

where Φ is the probability integral;

$h_0^2 = Q^2/\sigma^2$ is the signal/noise ratio;

Q^2 is the signal energy;

$\rho = \overline{T \cdot A_1(t) A_2(t)}/Q^2$ is the cross correlation coefficient.

For FT the equation for the signals have the form

$$\left.\begin{aligned} A_1(t) &= A_0 \cos(\omega_1 t + \varphi_0) \\ A_2(t) &= A_0 \cos(\omega_2 t + \varphi_0) \end{aligned}\right\} 0 \leqslant t \leqslant \tau_0.$$

For these signals

$$\rho = \frac{\sin[(\omega_1 - \omega_2)\tau_0]}{(\omega_1 - \omega_2)\tau_0}. \tag{I-35}$$

With a frequency separation equal to

$$\frac{\omega_1 - \omega_2}{2\pi} = \frac{k}{\tau_0},$$

where k is an integer, the condition $\rho = 0$ is satisfied and the potential noise factor of the FT is determined from equation

$$P_{errFT} = \frac{1}{2}[1 - \Phi(h_0)]. \tag{I-36}$$

For PT

$$\left.\begin{array}{l}A_1(t)=A_0\cos(\omega_0 t+\varphi_1)\\ A_2(t)=A_0\cos(\omega_0 t+\varphi_2)\end{array}\right\} 0\leqslant t\leqslant\tau_0.$$

For these signals /43

$$\rho=\cos(\varphi_1-\varphi_2). \quad \text{(I-37)}$$

In the case of 180° phase keying

$$P_{errPT}=\frac{1}{2}[1-\Phi(\sqrt{2}h_0)]. \quad \text{(I-38)}$$

Thus, the potential noise factor for phase modulation with phase shift by 180° coincides with the noise factor of the ideal binary system.

With pulse-time keying (PTT) the information is transmitted by means of the variation of the position of the pulse in time.

For this form of modulation, the signals are written in the form

$$\left.\begin{array}{l}A_1(t)=A_1\left(t+\frac{\Delta\tau}{2}\right)\cos(\omega_0 t+\varphi_0)\\ A_2(t)=A_2\left(t-\frac{\Delta\tau}{2}\right)\cos(\omega_0 t+\varphi_0)\end{array}\right| -\frac{\tau_p}{2}<t<\frac{\tau_p}{2},$$

where τ_p is the pulse duration.

Here

$$\rho=\frac{\overline{A_1\left(t+\frac{\Delta\tau}{2}\right)A_2\left(t-\frac{\Delta\tau}{2}\right)}}{Q^2}. \quad \text{(I-39)}$$

If the pulses do not overlap, then $\overline{A_1\left(t+\frac{\Delta\tau}{2}\right)A_2\left(t-\frac{\Delta\tau}{2}\right)}=0$ and

$$P_{errPTT}=\frac{1}{2}[1-\Phi(h_0)], \quad \text{(I-40)}$$

which coincides with the potential noise factor of FT.

Let us consider the case of transmission of orthogonal signals which can take m discrete values

$$A_1(t),\ A_2(t),\ \ldots,\ A_m(t).$$

This system of signals is termed orthogonal, if

$$\overline{A_i(t)\cdot A_j(t)} = 0 \text{ with } i \neq j$$

It is shown in reference 25 that the potential noise factor of the system with orthogonal signals is determined by equation /44

$$P_{err} = 1-\frac{1}{\sqrt{2\pi}}\int_{-\infty}^{+\infty}[1-v(\sqrt{2}\cdot h_0+y)]^{m-1}\, e^{-\frac{y^2}{2}}dy, \qquad (I-41)$$

where

$$v(x)=\int_x^{\infty} e^{-\frac{t^2}{2}}\,dt.$$

For $P_{err} \leq 0.1$ the following approximate equation (ref. 26) is valid with a high degree of accuracy

$$P_{err} \approx (m-1)[1-F(h_0)]. \qquad (I-42)$$

Reference 27 considers a telegraph system in which 32 orthogonal channels are transmitted. A definite signal corresponds to each letter in this system. Such signals can be, for example,

$$\left\{\begin{array}{l|l}
A_1(t)=A_0\cos(\omega_0 t) & 0\leqslant t\leqslant \tau_0, \\
A_2(t)=A_0\cos(\omega_0 t+\Delta\omega t) & \\
\cdots\cdots\cdots\cdots\cdots\cdots & \\
\cdots\cdots\cdots\cdots\cdots\cdots & \frac{\Delta\omega}{2\pi}=\frac{k}{\tau_0}, \\
\cdots\cdots\cdots\cdots\cdots\cdots & \\
A_{32}(t)=A_0\cos(\omega_0 t+31\Delta\omega t) & \text{k is an integer.}
\end{array}\right.$$

The amplitudes A_0 and durations τ_0 of the signals for all letters are identical, while the frequencies are different.

Figure I-13 presents the variations of the distortion probability P_{dis} with transmission of the letters by orthogonal signals with m = 32 and by binary signals with a five-digit code (FT).

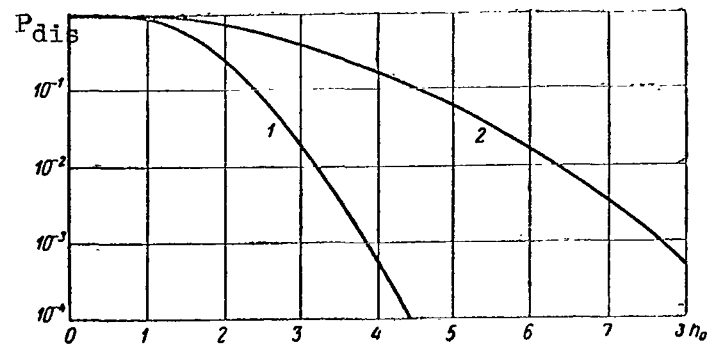


Figure I-13. Distortion probability for 32 signals. 1, orthogonal system; 2, binary system with five-digit code.

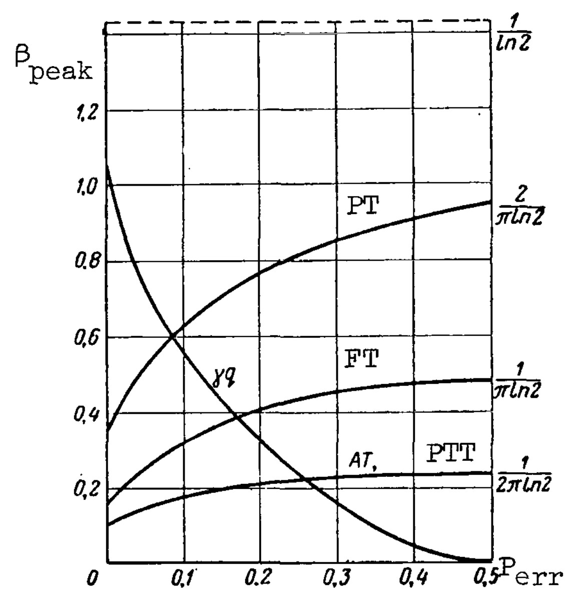


Figure I-14. Effectiveness of binary communication system.

Analysis of the curves shows that the orthogonal system with 32 signals provides a higher degree of noise immunity than the binary system with the five-digit code. With identical transmission rate, to provide the same error probability in the orthogonal system the required signal power is less by a factor of 3.5 than in the binary system.

References 28 and 29 carry out an evaluation of the β-efficiency, the η-efficiency and the γ-efficiency of the binary and multiple-state orthogonal systems. The results of the calculations for the binary systems are shown in figure I-14, where q denotes the quantity $q = F\tau_0$ (with optimal filtering q = 1).

The curves of η_2 for the various systems shown in figure I-15 are constructed for q = 1. /46

From the figures it follows that, from the point of view of the utilization of the power and the bandwidth, the PT is a more advanced system than the FT and the PTT. In the case of the multistate orthogonal systems, the utilization of the power and the frequency bandwidth for a given error probability improves with increase of m. The limiting value of β for the orthogonal system as $m \to \infty$ is equal to 1/2 ln 2.

The significant advantages of the multistate systems over the binary (high noise immunity, high efficiency) make it possible to consider that in the future the orthogonal multistate systems will find broad acceptance in space radio communication systems. At the present time the technical implementation of these systems is retarded, primarily, by the complexity of the equipment realizations and the difficulties in the selection of the family of orthogonal signals. As the result, we shall consider hereafter only the PT, FT and PTT /47 systems.

Phase telegraphy is the optimal coherent system for the transmission of binary signals. A receiver consisting of a coherent detector and an integrating device connected ahead of and behind the detector realizes the potential noise immunity. The practical implementation of such a receiver encounters difficulties, in particular, in obtaining the coherent voltage. Many circuits have been proposed in which the coherent voltage is obtained by suitable processing of the received signal.

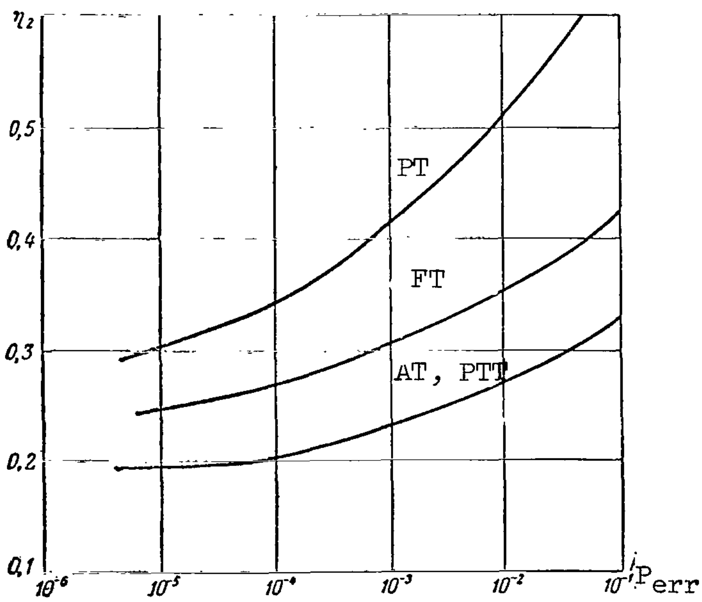


Figure I-15. η-effectiveness of binary system.

The theoretical and experimental investigations of these circuits have disclosed two essential difficulties: the tendency to reverse operation and the necessity for regulation of the circuit prior to operation and after interruptions of communication. A new method of phase telegraphy is the method of relative phase telegraphy (RPT) proposed by N. T. Petrovich (refs. 30 and 126). This method makes it possible to design a receiver which realizes the potential capabilities of phase keying, and in principle does not have the tendency to reverse operation. In the RPT system the information is transmitted by the relative value of the signal phase rather than by the absolute value: the phase of the nth signal is measured relative to the phase of the (n-1)th signal.

The noise rejection of the RPT system depends on the method of reception of the signals. Most promising are the methods of phase comparison and polarity comparison (ref 31).

The block diagram of the RPT receiver using the phase comparison method contains a signal delay circuit with a delay equal to the elementary train and a phase detector. In this case detection consists in the comparison of the phases of two neighboring signals. The ambiguity of the initial phases of the signals is identical and is eliminated during the comparison of the phases in the detector. With this method of reception the expression for the error probability takes the form

$$P'_{errRPT} = \frac{1}{2} e^{-h^2}. \qquad (I-43)$$

In the case of RPT reception using the polarity comparison method (coherent method of reception of the RPT signals), the received signal is first detected using the conventional coherent detector, and is then applied to the comparison circuit; here a comparison is made not of the phase, but of the polarity of the signals obtained at the detector output. For the calculation of the error probability we obtain expression (ref. 32) /48

$$P''_{errRPT} = \frac{1}{2}[1 - \Phi^2(\sqrt{2}h)] \qquad (I-44)$$

Curves plotted using equations (I-43) and (I-44) are shown in figure I-16. Analysis of the curves of figure I-16 shows that the polarity comparison method and the phase comparison method are approximately equivalent.

The computations of the noise factor of the RPT for both cases were made on the assumption that the frequency shift due to the various destabilizing factors Δf_{des} is small in comparison with the signal spectrum width. For the radio communication link with relay through the AES, whose operating frequencies according to preliminary estimates lie in the vicinity of 1,000-2,000 Mcps with operating rates of the order of 2,000 baud, this assumption is satisfied if the overall frequency instability does not exceed 10^{-6} (refs. 15, /49

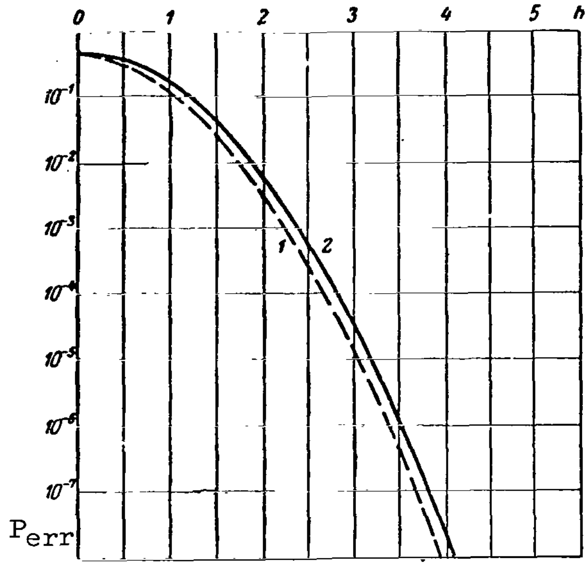


Figure I-16. Error probability in phase telegraphy systems. 1, RPT with coherent reception; 2, RPT with phase comparison reception.

121 and 124). This condition is quite severe in view of the fact that the frequency shifts due to the Doppler effect are large (of the order of 10^{-4} to 10^{-5}), and that it is difficult to maintain a high frequency stability (above 10^{-4}) on board the spacecraft over a long time period.

For this reason the given assumption $\Delta f_{des} \ll 2F_m$ for the case of the transmission of relatively narrow-band width information is usually not satisfied. With fulfillment of the reverse condition $\Delta f_{des} \gg 2F_m$ the receiver passband at the RF (ahead of the detector) must be considerably wider than is required for the passage of the signal. Since the power of the fluctuating noise at the receiver input is proportional to its passband width, the signal/noise ratio at the receiver input decreases with the increase of $\Delta f_{des}/2F_m$ and the reception of the signals is degraded.

Thus, we can draw the conclusion that the use of phase keying for the transmission of relatively narrow-band information over radio links with relaying via AES is justified only with a high degree of accuracy of the compensation for the Doppler shift of the frequency (on the order of 10^{-6}), and with the use on board the spacecraft and on the ground of high-stability frequency references (with stability of the order of 10^{-6})

Let us turn to the consideration of frequency modulation. Two methods of reception of frequency-modulated signals are used: using the instantaneous values of the envelope and using instantaneous values of the frequency.

In the first method the received signals are separated by two narrow-band filters and applied to amplitude detectors. The voltages from the detector outputs are compared with one another. The recording of the signal is performed depending on which of these voltages is the larger. The error probability in the FT system with identical a priori probabilities of the transmitted signals is equal to /50

$$P'_{errFT} = \frac{1}{2} e^{-\frac{h^2}{2}}. \tag{I-45}$$

For coherent detection with integration of the signal and the noise the error probability in the FT system is equal to

$$P''_{errFT} = \frac{1}{2}[1 - \Phi(h)]. \tag{I-46}$$

Thus, envelope reception with integration and coherent detection realizes the potential noise immunity of FT.

The circuit for the reception of the FT signals using the instantaneous value of the frequency consists of a limiter and a discriminator. An analytic expression for the error probability in this case has not yet been obtained.

Curve 2 in figure I-17 shows the variation of $P'''_{err_{FM}}$ with h obtained by numerical integration (ref. 33). Comparison of it with curve 1, plotted from equation (I-45), shows that reception using the instantaneous value of the /51 frequency is less noise-resistant than reception using the instantaneous value of the envelope.

The presence of the frequency shifts due to the Doppler effect and due to the various instabilities here again leads to a decrease of the noise rejection (in comparison with the potential). It is shown in reference 34 that with detection of the envelope in the case where the condition $N_m\left(2F_m/F\right) > 1$ is satisfied (where N_m is the signal/noise ratio at the output of the low frequency filter with passband F_m; F is the passband at the RF ahead of the detector) it makes no difference how the filtering is accomplished--at the low or high frequency. In this case it is advisable to use a linear detector. With the linear detector the signal/noise ratio at the output N_m is related as follows to the signal/noise ratio at the input of the detector N_{01}

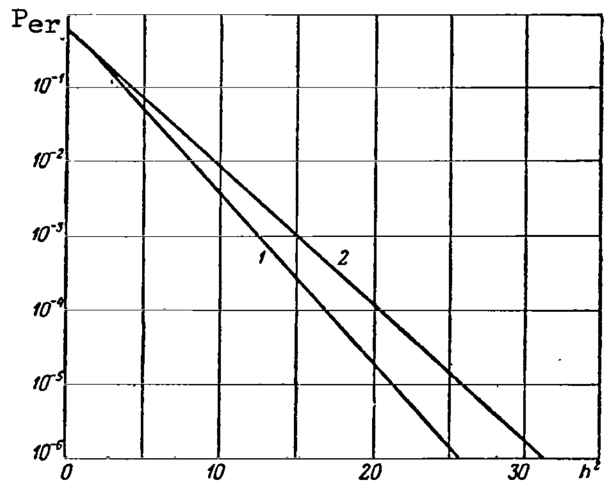


Figure I-17. Error probability in FT system with reception. 1, using instantaneous value of envelope; 2, using instantaneous value of frequency.

$$N_{0_1} \approx N_m \frac{2F_m}{F}. \qquad \text{(I-47)}$$

With satisfaction of the reverse inequality, it is advisable to use a square-law detector; in this case

$$N_{0_2} \approx 1{,}27 \left[N_m \frac{2F_m}{F} + \sqrt{\left(N_m \frac{2F_m}{F}\right)^2 + \frac{2}{3}\left(N_m \frac{2F_m}{F}\right)}\right], \qquad \text{(I-48)}$$

where the factor 1.27 characterizes the suppression of the signal in the limiter.

In the case of a weak signal and the use of the square-law detector, the actual FT receiver has a loss in comparison with the potential of a factor of N_{02}/N'_{01}, where N'_{01} is the required signal/noise ratio at the receiver input with linear filtering (for example, with a coherent detector). If the frequency shifts due to the various instabilities and the Doppler effect are large ($\Delta f_{des} \gg 2F_m$), then the square-law detector will give a very significant loss in comparison with the coherent detector. With $2F_m$ = 2 kcps, F = 200 kcps (which corresponds to an overall instability of $\pm 10^{-4}$ with operation at the frequency f = 1,000 Mcps), and with N_{m1} = 1, N_{m2} = 10, the losses will be /52 n_1 = 11.8 and n_2 = 4.75, respectively. If we consider that in the reception of FT signals the IF passband is selected equal to triple the channel filter passband, which is chosen as $\Delta f_{des} + 2F_m$, then the actual FM receiver gives losses of $n'_1 = 19.2$ and $n'_2 = 7.1$, respectively, in comparison with the potential receiver for the given value of instability $\pm 10^{-4}$.

Thus, effective application of FT for the retransmission of relatively narrow-band information (thousands of baud) over the space communication links, just as PT, requires compensation for the Doppler frequency shift and the use of highly stable frequency references.

Let us consider operation using pulse-time telegraphy (PTT). It was shown that the potential noise stability of PTT for the transmission of binary signals coincides with the noise stability potential of frequency keying (FT). Consequently, we can set $N_{mPTT} = N_{mFT}$. Let us assume that synchronous PTT is accomplished--the receiver is gated only at the time of possible arrival of the pulses. In the pulse method of operation the passband width at the low frequency $F_m = MF'_m$, where M is the duty cycle, F'_m is the operating speed, and for sufficiently large M (on the order of 1,000) the following condition is satisfied

$$N_m \frac{2F'_m M}{2F'_m M + \Delta f_{des}} > 1$$

even with large frequency shifts due to the Doppler effect and with various instabilities Δf_{des}. In this case it is advantageous to use a linear detector. The quantities N_{mPTT} and N_{OPTT} with the use of the linear detector and with a strong signal are connected by the following relation

$$N_{OPTT} = N_m \frac{2F'_m M}{2MF'_m + \Delta f_{des}} \qquad \text{(I-49)}$$

Consequently, this method of reception realizes a potential noise immunity of PTT which coincides with the potential noise immunity of FT. /53

This discussion leads to the following conclusions

(1) With relaying of radiotelegraph signals via AES (active and passive relaying), from the point of view of reduction of the power consumption in the radio link, it is advisable to make use of relative phase telegraphy (RPT) for the transmission, if the frequency shifts due to the Doppler effect and the various instabilities are compensated with a high degree of accuracy and the residual frequency instability does not exceed 10^{-6}-10^{-5}.

(2) With relaying of radiotelegraph signals via AES at relatively low speeds (tens of baud), when the signal spectrum width is much less than the residual frequency shifts due to the various instabilities and the Doppler effect, and also in the absence of compensation of such frequency shifts (even at high operating speeds--on the order of thousands of baud), it is advantageous to use the wide-band pulse methods of modulation, particularly PTT.

The further discussion will be made on the assumption that there is no compensation for the frequency shifts resulting from the Doppler effect and the various instabilities, and therefore we shall consider only the case of relaying signals with PTT. For definiteness, we assume $N_m = 22.5$. In this case the error probability will not exceed 10^{-5}, i.e., we consider the case of transmission of telegraph signals with high reliability. With account for post-detector filtering, the coefficient N_0F in the communication equation is written as $N_0F = 45 \cdot F'_m \cdot M$, where F'_m is the receiver passband after the detector; M is the duty cycle. For the case $F'_m = 1{,}000$ cps and $M = 1{,}000$, in the determination of the transmitter peak power $N_0F = 4.5 \cdot 10^7$ cps.

For the transmission of continuous information the widest usage has been made of amplitude modulation (AM), amplitude modulation with single or double sideband and suppressed carrier (SSB and DSB), frequency modulation (FM), and pulse-code modulation (PCM).

With AM the signal is written in the form /54

$$A(t) = A_0 [1 + dF(t)] \cos(\omega_0 t + \varphi_0),$$

where F(t) is the modulating function.

The gains B and B' are found from equations (21)

$$B = \frac{d^2 \cdot \nu}{k_p^2 + d^2}, \quad B' = \frac{d^2}{k_p^2 + d^2}, \tag{I-50}$$

where ν is equal to the ratio F/F_m;

d is the modulation factor;

k_p is the information peak factor.

With d = 1 the gain is maximal

$$B' = \frac{1}{k_p^2 + 1}.$$

If $k_p = \sqrt{2}$ (sinusoidal modulation), then $B' = 1/3$; if $k_p = 3$ (telephone modulation), then $B' = 1/10$.

For the DSB system

$$A(t) = dA_0F(t)\cos(\omega_0 t + \varphi_0)$$

and $B = \nu$, $B' = 1$. Consequently, the DSB system gain does not depend on the modulation factor or the information peak factor. With sinusoidal modulation and $d = 1$ the DSB system gain is twice that of the AM system.

For the SSB system with transmission of information using the upper frequency band, the signal will be equal to

$$A(t) = A_0 \sum_{i=i_1}^{i_2} \left[\lambda_{2i-1} \sqrt{2} \sin\left(\frac{2\pi i}{T} + \omega_0\right) t + \right. \\ \left. + \lambda_{2i} \sqrt{2} \cos\left(\frac{2\pi i}{T} + \omega_0\right) t \right], \tag{I-51}$$

where λ_i is a parameter defining the signal being transmitted.

Then $B = \nu$ and $B' = 1$, i.e., the SSB and DSB systems are the same with respect to noise immunity. /55

According to reference 27, the potential noise stability is realized with any noise level in the linear modulation systems (AM and DSB) and also in the SSB systems in the real receiver with coherent reception. With noncoherent reception this is valid only for a low noise level.

Frequency modulation (FM) is an example of the integral systems. In the case of FM the signal is written in the form

$$A(t) = A_0 \cos[\omega_0 t + \Delta\omega_m \Psi], \tag{I-52}$$

where $\Delta\omega_m$ is the frequency deviation and

$$\Psi = \int F(t)\,dt = \sum_{i=i_1}^{i_2} \left(-\frac{T\lambda_{2i-1}}{2\pi i} \sqrt{2} \cos\frac{2\pi i}{T} t + \right. \\ \left. + \frac{T\lambda_{2i}}{2\pi i} \sqrt{2} \sin\frac{2\pi i}{T} t \right).$$

For the gain B and the generalized gain B' we have

$$\left.\begin{aligned} B &= \frac{3}{k_p^2}\, m^2 \nu, \\ B' &= \frac{3}{k_p^2}\, m^2, \end{aligned}\right\} \tag{I-53}$$

where m is the modulation index, equal to $\Delta\omega_m/2\pi F_m$.

For the transmission of the same information in the case of FM operation, less power is required than in operation using AM by a factor of n, where $n = 3m^2 (1 + k_p^2)/k_p^2$.

The advantage obtained in the use of FM in comparison with AM is obtained primarily as the result of the expansion of the receiver passband. The frequency band occupied by the FM signal, with an accuracy sufficient for practical purposes, can be expressed by the simple equation (ref. 35)

$$F = 2F_m (1 + m). \tag{I-54}$$

If the power radiated by the transmitter is constant, then the higher noise stability of FM can be obtained only by means of increasing the modulation /56 index m. With increase of m there is an expansion of the frequency band occupied by the FM signal and an increase of the intensity of the fluctuating noise at the receiver input. Relation (I-53) is valid only with a high signal/noise ratio at the input of the FM receiver.

An approximate analysis of the case of high noise carried out by Kotel'nikov (ref. 27) has shown that with high noise the gain obtained with FM in comparison with AM is reduced, and this reduction is greater the larger m. Thus, FM is a threshold system. According to reference 21 the threshold value of the signal/noise ratio at the input of the FM receiver $N_{0_{thr}}$ varies approximately in the range from 1 to 2 with variation of the modulation index.

It is evident that with a constant signal power there is no sense in increasing the modulation index above some value m_{max} determined from the condition $N_0 = N_{0_{thr}}$. For the determination of m_{max}, from (I-53) and (I-54) we obtain equation

$$m^2 (m + 1) = \frac{N_m}{N_{0_{thr}}} \frac{k_p^2}{6}. \tag{I-55}$$

Thus, the optimal modulation index is higher, the higher the demands on the quality of the information (the quantity N_m) at the receiver output. With a modulation index determined by expression (I-55), the given value of N_m is achieved with minimal consumption of signal power.

The resulting threshold value $N_{0_{thr}} = 1\text{-}2$ is a limiting value and cannot be realized in real receivers. In real FM receivers the values of $N_{0_{thr}}$ obtained are considerably higher than those predicted by the theory for the potential receiver. From the experimental data (ref. 36) the threshold in the FM receiver with conventional reception occurs approximately with equality of the peak values of the signal and noise, which corresponds to $N_{0_{thr}} \approx 10$. According to other sources (ref. 37) $N_{0_{thr}} \approx 16$. By improving the reception methods we can improve the noise factor of the real FM receiver. With a /57 constant transmitter power the threshold is reached more quickly the higher the received modulation index m (as a result of broadening the receiver passband). In order to eliminate this effect we should reduce the receiver passband at the RF, which cannot be done with the conventional method of FM signal reception without the appearance of large nonlinear distortions. However, this reduction of the passband can be accomplished without increase of the nonlinear distortions of the signal, if we use the method of FM reception with frequency feedback, i.e., the method in which the resonant frequency of a tank circuit located ahead of the receiver frequency detector is automatically controlled by the voltage applied from the output of the frequency detector (ref. 38). Experiments show that with this method of reception we can reduce the threshold several times and can make it approach the limiting value (ref. 125).

In those cases when high fidelity of the reproduction of the transmitted information is required, operation using the FM method gives considerable advantage in comparison with the use of AM even with conventional reception methods. This can be illustrated by the following numerical example.

Let us consider the transmission of a television signal with low frequency spectrum width $F_m = 5$ Mcps. According to the IRCC (International Radio Consultative Committee) standards, for high-fidelity reproduction of a television signal a low-frequency signal/noise ratio $N_m = 10^4$, i.e., 40 db, is required. Let us take $k_p^2 = 2$. Taking $N_{0_{thr}} = 16$, we solve equation (I-55) relative to m and find $m_{max} = 5.6$. The FM signal spectrum width is equal to F = 66 Mcps. The required signal power at the input of the real FM receiver is equal to

$$P_{c_{FM}} = 1.06 \cdot 10^9 \; kT_{eff} \; (w),$$

where k is the Boltzmann constant, equal to $1.4 \cdot 10^{-27}$ w·sec/deg;

T_{eff} is the effective noise temperature at the receiver input, ^{o}K.

With operation using amplitude modulation with a value of the peak factor $k_p^2 = 2$, in order to provide $N_m = 10^4$ the required signal/noise ratio at the receiver input is equal to $N_0 = 1.5 \cdot 10^4$ and the signal power is /58

$$P_{c\,AM} = 1{,}5 \cdot 10^{11} kT_{eff}$$

The gain in power with the use of FM in comparison with AM is n = 140.

Let us make an estimate of what the potential FM signal receiver can do. In this case $N_{0thr} = 2$. Solving (I-55) relative to m, we find $m_{max} = 11.4$. The FM signal spectrum width is F = 124 Mcps. The required signal power at the receiver input $P_{c_{FM}} = 0.248 \cdot 10^9 kT_{eff}$. The power gain in comparison with AM is n' = 600-fold, and in comparison with the FM case with $m_{max} = 5.6$, n' = 4.2. Thus the use of the real FM receiver gives a power loss of 4.2.

Let us turn to the consideration of the pulse-code modulation (PCM). According to the theorem of Kotel'nikov, the continuous signal F(t) can be transmitted over a communication line with the required accuracy by means of the transmission of the individual instantaneous values of this signal

$$F(-2T_i),\ F(-T_i),\ F(0),\ F(T_i),\ F(2T_i). \tag{I-56}$$

taken for the instants of time separated from one another by the amount $T_i \le 1/2F_m$, where F_m is the highest frequency contained in the signal F(t). In the PCM system the signal is a sequence of code combinations which reflect the level-quantized values (I-56) of the transmitted information. The repetition period of the combinations is chosen equal to the period T_i of sampling of the instantaneous values of the signal F(t). Each code combination contains N elementary trains of identical duration. Considering that in the general case these trains can take k values, we see that by this method we can transmit $L = k^N$ levels of the signal F(t). Under the condition that these levels are equally probable, the transmission rate will be /59

$$V = p \log L = Np \log k, \tag{I-57}$$

where p is the sampling frequency.

With $p = 2F_m$ for the case of binary coding, which will be considered hereafter,

$$V_0 = 2NF_m. \tag{I-57'}$$

For the transmission of the trains of code combinations, we can make use of any of the methods of transmitting discrete information signals--AT, FT, PT (amplitude, frequency and phase keying). Thus, again here we see the advantage of PCM, just as in all systems based on discrete information, over the continuous transmission systems, since the systems with discrete information make it possible to use the more advanced modulation methods (relative phase keying with 180° phase shift, for example) and to use the more efficient methods of reception of weak signals (storage methods, correlation methods, etc.). The possibility of pulse regeneration with discrete transmission makes it possible to practically eliminate the accumulation of noise during multiple relays.

A characteristic of PCM, just as the other systems with quantizing of information, is that even with complete absence of noise in the channel the received information F(t) differs from the actual information $F_0(t)$ by the magnitude $\epsilon(t) = F(t)-F_0(t)$, where $\epsilon(t)$ is the "quantizing noise." Consequently, in the design of noise-resistant PCM systems it is necessary to take account of both the fluctuating noise and the quantizing noise.

The ratio of the signal power to the quantizing noise power at the outlet of the PCM system according to reference 21 is equal to

$$N_{m_k} = \frac{3L^2}{k_p^2} = \frac{3 \cdot 2^{2N}}{k_p^2}. \tag{I-58}$$

From this we see that by selecting a sufficiently large number of quantizing levels L (or a large number of pulses in the combination N) we can obtain any desired ratio of the signal power to the quantizing noise power. However, we cannot increase the noise stability of the PCM without limit in this way. /60 It can be shown that in this case the fluctuation noise increases; its presence leads to the creation of false pulses in the channel or to suppression of the signal pulses. Actually, increase of the number of pulses N leads to an expansion of the signal spectrum and to a broadening of the required receiver passband. The repetition frequency of the code combinations must be no less than $2F_m$ for one channel and $2F_m n$ for an n-channel system. Consequently, the maximal pulse repetition frequency cannot be less than $2NnF_m$ or the duration of the pulse τ_0 cannot be greater than $1/2N \cdot n \cdot F_m$. As is known, for the transmission of a radio pulse of duration τ_0 we need a frequency band

$$F=\frac{2a_2}{\tau_0},$$

where a_2 is a coefficient which depends on the permissible distortion of the pulse shape.

Evidently, since $a_2 \geq 1$,

$$F \geqslant 4NnF_m. \tag{I-59}$$

We see that F is directly proportional to N. The broadening of the receiver passband leads to an increase of the fluctuating noise power at the input and output of the receiver. The ratio of the signal to the fluctuating noise at the output with PCM is determined from the equation

$$N_m=\frac{3}{4}\cdot\frac{(L-1)^2\cdot p}{k_p^2\,(L^2-1)\cdot F_m\cdot[1-\Phi(\alpha)]}. \tag{I-60}$$

This equation is valid for any form of modulation of the carrier frequency. Only the value of α depends on the form of modulation in this equation.

In the case of amplitude modulation (PCM-AM)

$$\alpha^2=\frac{1}{2}h_0^2$$

and the signal/noise ratio at the receiver input is

$$N_0=\frac{P}{\sigma^2\cdot F}=\frac{2\alpha^2 NnP}{F},$$

and the system gains B, B' for $k_p^2 = 2$ are found from equations (ref. 21) /61

$$\left.\begin{aligned} B_{AM}&=\frac{3}{8}\cdot\frac{(L-1)^2}{(L^2-1)}\cdot\frac{\nu}{Nnh_0^2\left[1-\Phi\left(\frac{h_0}{\sqrt{2}}\right)\right]},\\ B'_{AM}&=\frac{B_{AM}}{\nu}.\end{aligned}\right\} \tag{I-61}$$

For the PCM-FM system (frequency modulation of the carrier) $\alpha^2 = h_0^2$ and

$$
\left.\begin{aligned}
B_{FM} &= \frac{3}{8}\cdot\frac{(L-1)^2}{(L^2-1)}\cdot\frac{\nu p\tau_0}{h_0^2[1-\Phi(h_0)]},\\
B'_{FM} &= \frac{B_{FM}}{\nu}.
\end{aligned}\right\} \qquad \text{(I-62)}
$$

For the PCM-PM system (phase modulation of the carrier) $\alpha^2 = 2h_0^2$ and

$$
\left.\begin{aligned}
B_{PM} &= \frac{3}{8}\cdot\frac{(L-1)^2}{(L^2-1)}\cdot\frac{\nu p\tau_0}{h_0^2[1-\Phi(\sqrt{2}h_0)]},\\
B'_{PM} &= \frac{B_{PM}}{\nu}.
\end{aligned}\right\} \qquad \text{(I-63)}
$$

For the case $\alpha^2 \gg 1$ the equations for the PCM gain take the form

$$B_{AM} \approx \frac{2}{k_p^2\sqrt{N_0}}e^{\frac{N_0}{2}}, \qquad \text{(I-61')}$$

$$B_{FM} \approx \frac{2\sqrt{2}}{k_p^2\sqrt{N_0}}e^{N_0}, \qquad \text{(I-62')}$$

$$B_{PM} \approx \frac{4}{k_p^2\sqrt{N_0}}e^{2N_0} \qquad \text{(I-63')}$$

PCM is a threshold system. In the systems which are close to optimal the threshold occurs when with a decrease of N_0 the fluctuation noise begins to exceed the quantization noise.

Let us determine the threshold value of the signal/noise ratio at the input of the potential PCM receiver for the case of speech transmission /62

$$
\begin{aligned}
&k_p^2 = 10,\ N = 10,\ n = 1,\ F_m = 3 \text{ kcps},\\
&N_{m_1} = 10^4 (40 \text{ db}),\ N_{m_2} = 10^6 (60 \text{ db}),\\
&N_{m_3} = 10^8 (80 \text{ db}).
\end{aligned}
$$

From (I-58) we find the number N of pulses in the code combination $N_1 = 8$, $N_2 = 11$, $N_3 = 15$.

Equating $N_{m_k} = 2\cdot10^4;\ 2\cdot10^6;\ 2\cdot10^8$ and considering that $N_m = BN_0$, we find $N_{0_{thr}}$ for the different forms of modulation of the carrier with operation using PCM

$$N_{0_{thr}} \approx 20.5;\ 29.5;\ 38 \text{ (PCM-AM)};$$

$$N_{0_{thr}} \approx 10;\ 14.5;\ 19 \text{ (PCM-FM)};$$

$$N_{0_{thr}} \approx 5;\ 7;\ 9.5 \text{ (PCM-PM)}.$$

We see that the lowest threshold value (highest noise immunity) is that of the PCM-PM system. It is of interest to make a comparison of the potential noise stability of PCM and FM for the case of transmission of the same information--speech--with $F_m = 3$ kcps, $k_p^2 = 10$ and $N_m = 10^4;\ 10^6;\ 10^8$.

Let us consider the case of PCM-PM when the code combination consists of pulses modulated by means of 180° phase shift. In this case $F = 4N \cdot F_m = 96$, 132, 180 kcps, respectively, for $N_m = 10^4;\ 10^6;\ 10^8$. The power required at the input of the PCM-PM receiver is

$$P_{c_{PCM}} = 4.8 \cdot 10^5 kT_{eff};\ 9.25 \cdot 10^5 kT_{eff};\ 17.1 \cdot 10^5 kT_{eff}.$$

For the case $k_p^2 = 10$ and $N_{0_{thr}} = 2$ (we consider the potential FM receiver) we obtain the following values of the modulation index: $m_{max} = 20.2,\ 94,\ 436$. The RF spectrum width of the FM signal will be $F = 127$, 570, 2,620 kcps. The power required at the FM receiver input is

$$P_{c_{FM}} = 2.5 \cdot 10^5 kT_{eff};\ 11.4 \cdot 10^5 kT_{eff};\ 52.4 \cdot 10^5 kT_{eff}.$$

Thus, with large N_m, as we increase the requirements on the accuracy /63 of the reproduction of the information, there is an increase in the advantage of PCM in comparison with FM in respect to the required signal power. Here we must also take account of the fact that in the real PCM receiver it is easier to realize a noise stability close to the potential value than it is in the real FM receiver. We remarked above that with the conventional method of FM signal reception $N_{0_{thr}} \approx 16$. In this case we obtain the following values of the modulation index, $m_{max} = 10.1,\ 47,\ 218$. The RF spectrum width of the FM signal will be $F = 66.5$ kcps, 288 kcps, 1,310 kcps and the power required at the input of the real FM receiver is

$$P'_{c_{FM}} = 10.6 \cdot 10^5 kT;\ 46 \cdot 10^5 kT;\ 210 \cdot 10^5 kT\ [W].$$

In the case of PCM-RPT with reception using the polarity comparison method and with reception using the phase comparison method, the potential noise immunity of PCM-FM is practically fully realized, since the threshold signal/noise ratio at the input of the PCM receiver is quite high. As a result of the fact that the spectrum width of the PCM signal is quite broad and increases with increase of the demands on the fidelity of reproduction, the reduction of the real noise stability of the PCM relative to the potential value as the result of the frequency shifts because of the various instabilities and the Doppler shift will be small even with the transmission of relatively narrow-band infor- tion (for example, telephony with $F_m = 3$ kcps. This implies that at the input to the real PCM receiver there is required practically the same power as at the input of the potential receiver. Consequently, the PCM-RPT receiver for speech communications gives a gain in power, in comparison with the FM receiver, for the real receivers of n = 2.2, 5, 12.3, respectively, for $N_m = 10^4$ (40 db), 10^6 (60 db), 10^8 (80 db).

The value of the required signal power at the receiver input for the transmission of the same information characterizes the β-efficiency of the different forms of modulation--the most important communication system criterion with the existence of limitations on transmitter power. Thus, on the basis of these numerical calculations, we can conclude that the PCM-RPT system is, from /64 the point of view of providing the highest β-efficiency, the most advanced of all the presently existing systems for the transmission of information with high fidelity of reproduction of the message.

With a relaxing of the requirements of the quality of the reproduction of the information (reduction of N_m), the power gain of the PCM system in comparison with FM diminishes somewhat if we consider that the threshold signal/noise ratio for FM decreases with a reduction of the modulation index m. With $N_m = 10^3$ the power capabilities of PCM and FM are about equal. In the region $10^2 < N_m < 10^3$ the advantage of FM (or PCM) over AM is retained. For transmission of telephony signals with $k_p^2 = 10$ and $N_m = 10^3$ (high fidelity reproduction) the gain of FM in comparison with AM is 100; with $N_m = 10^2$ (medium fidelity reproduction) the gain of FM compared with AM is 20. With medium and low fidelity of reproduction ($N_m \leq 10^2$) of the information with a high value of k_p^2 (telephony, some forms of television images), such forms of modulation as SSB and DSB become competitive with FM also. For the transmission of telephone signals with $k_p^2 = 10$ and $N_m = 100$, SSB (DSB) the advantage of FM over SSB (DSB) with respect to power is only a factor of 2; with $N_m = 20$ the use of SSB now has a 2-fold advantage over FM.

In the space radio communication links, as a result of the rapid variation of the location of the corresponding stations and the associated large frequency shift (Doppler effect), and also the problem of maintaining a high frequency stability over a long period of time (on the order of 10^{-7} to 10^{-8}), there arise difficulties in obtaining a reference voltage for the coherent detector when using SSB and DSB reception. For this reason the effective utilization of those advantages of these forms of modulation such as the absence of the threshold for any noise level with coherent reception, the good utilization of the transmitter power and the energy advantage over FM with small N_m, form at the present time a quite complex problem.

Thus, of all methods considered for the transmission of continuous information the most promising from the point of view of application for space radio communication are FM and PCM-RPT. With requirement for high fidelity of reproduction of information ($N_m > 10^3$) the use of PCM-FM gives a power advantage in /65
comparison with FM which increases with increase of N_m. In the case of PCM, with increase of N_m the γ-efficiency (frequency band use factor, equal to the ratio of the rate of information transmission V to the receiver band-pass width at the radio frequency F) decreases slower than the γ-efficiency of FM, which is still another advantage of PCM over FM. With very high requirements of the quality of the information transmission, the advantages of PCM are unquestionable: the β-efficiency, γ-efficiency and η-efficiency of PCM are higher than for any continuous information transmission system known at the present time, and therefore pulse-code modulation should be used for the space radio communication links in cases with $N_m \geq 10^5$. With reduction of the demands on the accuracy of the reproduction ($N_m \leq 10^4$), PCM is comparable with FM with respect to criteria such as noise rejection and efficiency. In these cases the use of communication systems with FM is advisable in view of the greater simplicity of implementation (ref. 125).

We shall perform the calculations of the power requirements of a radio communication link with relaying through AES for the case of transmission of telephony (F_m = 3 kcps) and television (F_m = 5 Mcps). According to the IRCC standards for high quality reproduction of telephony and television, the signal/noise ratio N_m at the output of the receiver must not be less than 40 db.

According to the discussion above, in this case it is advisable to use the FM method. Therefore, we shall consider only the frequency modulation method. We previously made a comparative evaluation of FM and AM in the case $k_p^2 = 2$ for the transmission of a television signal, i.e., the information transmitted over the channel was a sinusoid. The peak factor of the real television signal is somewhat greater than $\sqrt{2}$. Reference 39 presents data on the amplitude distribution of the television signal. The evaluations of the peak factor of television information made on the basis of these data show that the most probable value is $k_p = 2.5$.

With the use of the conventional FM receiver $N_{0_{thr}} = 16$ and, conse- /66 quently, $m_{max} = 8.4$; if, however, we use a FM receiver with frequency feedback, then in the best case $N_{0_{thr}} = 2$, and then $m_{max} = 17$. Calculations show that the FM receiver with frequency feedback (with a "tracking filter") gives a maximal power gain of a factor of 4 in comparison with the conventional FM receiver. For this reason we shall not carry out power calculations separately for each of them, but shall compute only the case with reception using the conventional FM receiver. The FM signal spectrum width with $m = 8.4$ and $F_m = 5$ Mcps at the RF is equal approximately to $F = 94$ Mcps. In this case we can neglect the frequency shift due to the various instabilities and the Doppler effect, and the coefficient N_0F in the communication equation for all the radio links is constant and equal to $16 \cdot 94 \cdot 10^6$ cps $= 1.5 \cdot 10^9$ cps. For the transmission of speech with $k_p^2 = 10$ and use of the conventional FM receiver with $N_{0_{thr}} = 16$, the value of the modulation index will be $m_{max} = 10.1$, while with the use of a receiver with $N_{0_{thr}} = 2$ the modulation index $m_{max} = 20.2$. The utilization of a FM receiver with frequency feedback in place of the conventional FM receiver can give in the best case a reduction of the power consumption in the communication link by a factor of 4.5.

We shall also carry out the power consumption analysis for the radio link with a single telephone channel only for the case of reception using the conventional FM receiver, but now we cannot neglect the frequency shift resulting from various instabilities and the Doppler effect, and the factor $N_0 \cdot F$ in the communication equation is written in this case as

$$N_0F = 16\,[6.7 \cdot 10^4 + \Delta f_{des} + \Delta f_D]\ \text{cps},$$

where Δf_{des} is the frequency shift resulting from various instabilities;

Δf_D is the frequency shift resulting from the Doppler effect.

We must keep in mind that in the case of active relaying it is not economically sound to use the AES for the relaying of a single telephone or telegraph channel. The commercial communication must of necessity be multi- /67 channel. In this case multiple channel capability can be achieved in the following different ways.

(1) Direct connection through the AES of the ith correspondent with the kth, with quite a large number of communicating correspondents operating on a single frequency (common channel).

Since the various pairs of correspondents can initiate communication at any instant of time, even in the case of the pulse methods of operation, it is impossible to use time separation of channels. Here use can be made of channel separation (discrimination on Earth of the desired information from the mixture of signals from the different correspondents) on the basis of signal shape. Several systems of this sort have been described in the literature, for example the RADA system (Random Access and Discrete Address, i.e., a system with random selection of subscribers and discrete addresses) for the transmission of speech, and the digital systems with noise-like signals (refs. 40, 41 and 42).

In the RADA system delta modulation (DM) is used to convert the speech signals from analog form into sequences of pulses. Each of these pulses is transmitted by combinations of three or four short radio pulses. The variation of the time interval between pulses and the frequency selection of the trains give several thousand time-frequency address combinations, which makes it possible for several thousand correspondents to use a single common channel.

Schematically the broad-band communication systems with noise-like signals can be represented in the following form. At the transmitting end the information is first converted into a narrow-band signal by application of one of the known methods of modulation or keying. Then this signal is multiplied together with the "noise," which is generated with the aid of a special generator delivering a sequence of pulses having prescribed statistical characteristics. The signal thus obtained will be noise-like and with reception on a conventional receiver is sensed as noise. At the receiving end there is a noise-like-signal generator identical to that at the transmitting end. By multiplying the reference signal and the mixture arriving at the receiver /68 input, consisting of the signal (or signals from several correspondents) and ordinary fluctuating noise, we restore the narrow-band signal, which is then detected by the usual method.

To be specific in our consideration of the noise-like systems, we shall present one of the versions of the design of the block diagram of the Rake type single-band system (ref. 41, fig. I-18). The system block diagram consists of the receiver, two correlators, two reference generators (for the two forms of information--for the "mark" signal and for the "space" signal), a comparison circuit and an indication circuit. Each correlator consists of an amplifier, integrator and envelope detector. With arrival of the "mark" signal at the input of correlator 1, the reference and received signals coincide and the correlation function has a maximum. Consequently, at the output of this correlator there will be a maximal voltage. Similarly, with the transmission of the "space" signal the maximal voltage will be on the output of correlator 2. It is evident that if each pair of correspondents uses different pairs of /69 forms of noise-like signals and if these pairs of forms are approximately orthogonal, operation is possible without mutual interference by many pairs of correspondents using a single common channel.

In view of the inadequate study which has been made of the noise-like systems, we shall not make a power requirement analysis of the communication channel using form selection.

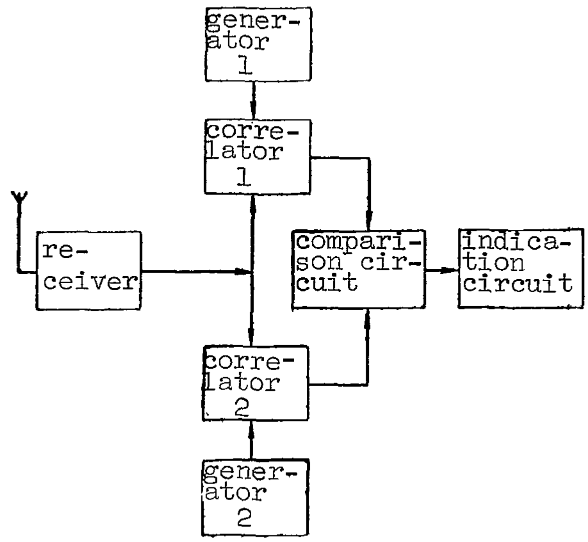


Figure I-18. Simplified version of block diagram of Rake type system with single ray.

(2) Direct communication via AES of the ith correspondent with the kth by reserving for each line i-k one channel on the repeater. Frequency separation of the channels is used.

In view of the severe limitations on weight and size of the repeaters on board the AES and the requirements for their high reliability, the number of lines i-k (and corresponding number of channels on the AES) must not be large. For this reason, this method of implementing multiple channelization in the near future will apparently find application only in lines for intergovernmental communication, official communication lines, etc.

In this case we shall make the analysis of the power requirement of the repeater transmitter on board the AES for a single channel. Having available a definite amount of dc power and knowing the radiation power of the onboard transmitter per channel, we can determine the potential possible number of channels (and correspondingly the number of pairs of correspondents i-k) of the repeater on board the AES. We shall make the calculation for the case of relaying of telegraph signals with pulse-time keying and telephone signals with FM.

(3) Relaying via the AES of a complex signal formed from the signals of many ground correspondents at some central transmitting point A_i (communication of correspondents with the central station is accomplished by conventional radio facilities or by cable), and reception of this complex signal either by the individual correspondents who separate from the complex signal the information addressed to them, or by a central receiving station A_k where decoding of the

individual messages and distribution to the addressees is accomplished by conventional radio equipment (using radio relay lines, for example) or by land ...

In this case, use is made of modulation systems with subcarrier frequencies. Here the number of subcarriers is chosen equal to the number of channels. /70

On the basis of the results of reference 21, we shall make a comparative evaluation of the noise immunity and the efficiency of the systems with subcarriers which have received the widest application in practice: FM-SSB, SSB-FM, FM-FM and SSB-SSB.[1]

The equations for finding the gain B will have the form

FM-SSB

$$B_k = \frac{3}{k_p^2} \cdot \frac{m_k^2 \cdot \nu}{k_f^2}, \tag{I-64}$$

SSB-FM

$$B_k = \frac{1}{k_p^2} \cdot \frac{\nu \Delta f_m^2}{k_f^2}, \frac{1}{f_k^2 + f_k F_{m_k} + \frac{1}{3} F_{m_k}}, \tag{I-65}$$

FM-FM

$$B_k = \frac{3}{2k_p^2} \cdot \frac{\nu \Delta f_k^2 \, \Delta f_m^2}{k_f^2 \, f_k^2 \, F_{m_k}^2} \tag{I-66}$$

SSB-SSB

$$B_k = \frac{\nu}{k_f^2}. \tag{I-67}$$

---

[1] The first letters give the abbreviated name of the method of modulation used on the subcarrier frequencies, and the second letters give the name of the modulation method used on the carrier frequency. For example, FM-SSB denotes that the subcarrier is frequency modulated, and that SSB modulation is used on the carrier frequency. We note that SSB-SSB is equivalent to the usual SSB.

In these expressions

$$k_p \approx n,$$

where n is the number of channels;

F_{m_k} is the passband of the kth channel for low frequency;

F is the overall passband width of the multiple channel system at the radio frequency;

f_k is the mid-frequency of the kth channel, equal to the subcarrier frequency;

Δf_k is the frequency deviation in the kth channel; /71

Δf_m is the overall frequency deviation;

m_k is the modulation index of the subcarrier frequency of the kth channel;

$$\nu = \frac{F}{F_{m_k}}.$$

We shall make the comparison for the case of the transmission of multichannel telephony with $k_p^2 = 10$, $n = 100$ and $F_{m_k} = 3$ kcps for the upper telephone channel (worst case). We assume that $N_{m_{100}} = 100$ (20 db).

The equations for B are valid for $N_0 > N_{0_{thr}}$. The noise stability threshold in the general case will hold for both modulation stages.

For the SSB-FM system the threshold is always determined by the carrier modulation system, and for the conventional FM reception method is equal to $N_{0_{thr}} = 16$. For the FM-FM system with $n = 100$ and $\Delta f_m / f_n > 3$ the threshold is determined by the carrier modulation system and also is equal to $N_{0_{thr}} = 16$. In the FM-SSB system the threshold is $N_{0_{thr}} = 16$. There is no threshold for the SSB-SSB system.

Calculations show that to obtain $N_m = 100$ in the upper telephone channel the minimal required power at the input of the multichannel receiver with the use of the various methods of modulation with subcarrier frequencies must have the following values.

For operation using FM-FM

$$P_{c_1} = 2\cdot10^8 kT_{eff},\ F_1 = 12 \text{ Mcps},$$

where F_1 is the receiver RF passband width; the carrier modulation index m_1 and the subcarrier m_2 are equal to $m_1 = m_2 = 3.15$.

For operation using SSB-FM

$$P_{c_2} = 1.4\cdot10^8 kT_{eff},\ F_2 = 9 \text{ Mcps}.$$

The carrier frequency modulation index is 14.5.

For operation using FM-SSB

$$P_{c_3} = 6.6\cdot10^8 kT_{eff},\ F_3 = 2.4 \text{ Mcps}.$$

For operation using SSB-SSB /72

$$P_{c_4} = 3\cdot10^9 kT_{eff},\ F_4 = 0.3 \text{ Mcps}.$$

We see that the use of SSB-FM gives a power gain of 1.4 in comparison with FM-FM, 4.7 in comparison with FM-SSB, and 22 in comparison with SSB-SSB.

Thus, in the case of relaying via AES of multichannel analog information, in the selection of the modulation method preference must be given to the SSB-FM system as having the highest β-efficiency. This fact is well known, and the SSB-FM system has found wide application in radio relay communication links.

We remarked that the use of SSB on the space communication links encounters several technical difficulties, resulting from the presence of large and time-varying shifts of the frequency because of the Doppler effect. It is evident that this remark does not apply to the SSB-FM system, since here single-band modulation is used on the subcarrier frequencies. For example, with n = 100 and F_m = 3 kcps the highest subcarrier frequency does not exceed 600 kcps even with a large margin r for the separation filtering of the channels (r = 2). With the most pessimistic assumptions on the stability of the onboard reference frequency, the overall frequency instability coefficient, taking account of the effect of the various instabilities and the Doppler effect, will apparently not be more than 10^{-4} and the shift of the frequency in each channel will not exceed 60 cps.

Just as in the case of the single-channel systems, the use on the space communication radio links of the multichannel systems with discretization of the information looks promising--the systems with pulse-code modulation (PCM). Here the introduction of additional subcarrier frequencies is not mandatory. It was shown earlier that of all the PCM systems, PCM-RPT has the best performance (noise rejection and efficiency). Calculations show that with use of this method, for n = 100, N_m = 100 the power required at the receiver output is $P_{c_5} = 1.5 \cdot 10^7 kT_{eff}$. The number of pulses in the code combination is N = 5, the required PCM receiver passband width is F_5 = 6 Mcps.

We can see that the use of PCM-RPT gives a power gain of a factor of 10 in comparison with SSB-FM. However, we should note that this comparison was somewhat unjustified, since in the SSB-FM system N_m = 100 only in the upper telephone channel, and with reduction of the channel number the quality of the reproduction improves; in the PCM-RPT system N_m = 100 for all channels. /73

Computations show that in the center telephone channel (n' = 50) N'_m = 500. In the case n = 100 and N_m = 500 the power required at the inlet of the PCM-RPT receiver is $P'_{c_5} = 2.5 \cdot 10^7 kT_{eff}$, and it must have a passband of 7.2 Mcps (the number of pulses in the code combination was taken as N = 6). Thus, even with comparison at the center telephony channel (n' = 50) the PCM receiver gives a gain of a factor of 5.5 relative to SSB-FM (using a conventional FM receiver).

This comparative analysis of the different multichannel communication systems makes it possible to draw the conclusion that the most promising from the point of view of suitability for the space radio communication links, where the primary index of the system is the β-efficiency, is the system of pulse-code modulation with the use of relative phase telegraphy (PCM-RPT); it is followed by the widely used SSB-FM system in radio relay communication links. All other multichannel systems are considerably inferior to these two in regard to the required power.

In view of the relatively complex technical implementation of the transmitting (in comparison with SSB-FM), and particularly the receiving devices of the PCM system with a large number of channels, it is apparent that the SSB-FM system will find wider use in the immediate future. For this reason we shall make the power requirement calculations for this system.

We showed previously that with the transmission of television signals with $N_m = 10^4$, F_m = 5 Mcps, N_0 = 16 the receiver RF passband must be 94 Mcps. With the transmission of telephony (n = 100) with N_m = 100 for the upper telephone channel and F_m = 3 kcps, the minimal receiver passband is equal to about 9 Mcps. Thus, the transmission of the television signal with these parameters using the FM method is equivalent, from the power requirement point of view, to the transmission using SSB-FM of the complex signal consisting of about 1,000 telephone signals (n = 1,000), with N_m = 100 for the upper telephone channel and F = 3 kcps. Therefore, we shall make the power requirement calculations only for the transmission of a television signal using FM, and note that one television channel with FM is equivalent to a certain number of telephone channels using SSB-FM. /74

Effective Noise Temperature at the Receiver Input. The effective noise temperature at the receiver input is determined by the intrinsic noise of the receiver (equivalent noise temperature T_0) and by the external noise (equivalent noise temperature T_{ex}). The noise temperature T_0 of the receiver corresponds to that temperature of a real resistor which can be converted into the noise power delivered to the active load of the real receiver. The cause of the appearance of the internal noise of the receiver are the electrical fluctuations in its various components and elements: resistors, condensers, coils and tubes.

All particles of matter, including the electrically charged particles, are in a state of thermal motion. The velocities of the movement of the free electrons vary randomly in magnitude and direction as the result of their collisions with one another and with other particles. The random motion of the charged particles in a conductor is equivalent to an electrical current which changes its magnitude and direction in accordance with a very complex law. This fluctuation current, passing through a conductor with resistance R, creates on its terminals a fluctuation voltage whose effective value is determined by the equation (ref. 43)

$$U_n^2 = 4kTR\Delta f, \tag{I-68}$$

where k is the Boltzmann constant;

T is the absolute temperature;

Δf is the frequency band.

One of the causes of the occurrence of noise in tubes is the shot effect.

The internal noise of the receiver is characterized by the noise coefficient s, which is the quantity indicating how many times in comparison with room temperature we must increase the absolute temperature of the conventional wire resistor equal to the input resistance R_{in}, in order that the noise current caused by this resistor will be equal to the input noise current. Usually the receiver noise coefficient is measured with an equivalent antenna connected to the receiver input. The equivalent self-noise temperature of the receiver is related with the noise coefficient thus determined by the equation $T_0 = (s'-1) \cdot 300^{\circ}K$. /75

It is known that the input stages of the receiver are the "noisiest."

Thus, to reduce the power consumption on the radio link we make use of low-noise input devices (amplifiers or mixers). Using triode amplifiers and traveling wave tubes as the input devices, we cannot obtain a noise coefficient lower than 6 db ($T_{0_1} = 900^{\circ}K$) at 1,000 Mcps, 8 db ($T_{0_2} = 1,600^{\circ}K$) at 2,500 Mcps and 10 db ($T_{0_3} = 2,700^{\circ}K$) at 10,000 Mcps (ref. 44).

With the use of triode amplifiers at the high frequencies (above 100 Mcps) the receiver sensitivity was limited by the self-noise, and the external noise was not usually taken into account. With the appearance of the low-noise receiving devices the situation was altered. It was found that in certain cases the receiver sensitivity was limited not by the receiver self-noise, but rather by the external noise. Studies showed that there are various types of external noise--noise from the Galaxy, discrete radio sources, Sun, planets, Earth's atmosphere, terrestrial, etc. /78

Before starting the analysis of the noises of various origin, let us introduce some characteristics of the radiation of cosmic objects.

According to reference 49, the radiation of any noise source is characterized by the brightness I and the radio emission flux P. By the flux of radio emission P, we mean the energy radiated by the source in a unit frequency interval, passing in unit time through a unit area in the direction normal to this area. The unit of measurement of P is W/m^2 cps. Brightness I characterizes the distribution of the intensity of the radio emission over the source.

The radio emission flux P and brightness I are connected by the relation

$$P=\int_{\Omega_e} I d\Omega, \quad (I-69)$$

where Ω_e is the solid angle from which the radiation is received.

The brightness of the radio emission is expressed usually in terms of the brightness temperature T_e, which is defined as the temperature of a blackbody, having at the given frequency in the given direction the same brightness as the source in question. In the RF band the radiation of a blackbody is determined by the Rayleigh-Jeans equation, according to which brightness I is associated with the brightness temperature T_e by the following relation

$$I=\frac{2kT_e}{\lambda^2}, \quad (I-70)$$

where k is the Boltzmann constant; /79

λ is the wavelength.

The radio emission flux is customarily expressed in terms of the effective temperature T_i, which is defined as the temperature of a blackbody having the angular dimensions of the source and radiating at the given frequency the same energy flux as the source in question.

According to this definition

$$P = \frac{2kT_i\Omega_e}{\lambda^2}. \tag{I-71}$$

The following relation between the effective and the brightness temperatures is obvious

$$T_i = \frac{1}{\Omega_e}\int_{4\pi} T_e(\varphi, \theta)\, d\Omega, \tag{I-72}$$

where it is considered that the brightness temperature T_e in the general case, just as brightness I, depends on the coordinates of the radiating region φ, θ.

With reception of nonpolarized radio emission, the power received in the frequency interval Δf from the elementary body angle $d\Omega$ and delivered to a load matched with the antenna is equal to

$$dP_c = \frac{1}{2} IA\, df\, d\Omega = \frac{kT_e(\varphi, \theta)}{\lambda^2} A\, df\, d\Omega, \tag{I-73}$$

where A is the effective antenna area, associated with the directivity coefficient of the antenna $G(\varphi, \theta)$ by the relation

$$A(\varphi, \theta) = \frac{\lambda^2}{4\pi} G(\varphi, \theta). \tag{I-74}$$

The total power P_c delivered by the antenna to the matched load in the frequency band df is equal to

$$P_c = \frac{k\Delta f}{4\pi}\int_{4\pi} T_e(\varphi, \theta) G(\varphi, \theta)\, d\Omega. \tag{I-75}$$

In our case, when we are comparing noises from various sorts of sources, it is convenient to make use of the concept of the equivalent source temperature T_a, referenced to the antenna. The equivalent source temperature, referenced to the antenna, is the temperature of a resistor equal to the input resistance of the antenna which delivers to the receiver load the same noise power as the source in question. The noise power delivered into a matched load by a resistor at temperature T_a is equal, as is known, to $P_c = kT_a\Delta f$. Equating this value of the power and the value expressed by equation (I-75), we obtain the following relation for the equivalent source temperature, referenced to the antenna /80

$$T_a = \frac{\int\limits_{4\pi} T_e(\varphi, \theta) G(\varphi, \theta) d\Omega}{\int\limits_{4\pi} G(\varphi, \theta) d\Omega}. \tag{I-76}$$

We introduce the concept of the effective solid angle of the antenna radiation pattern

$$\Omega_a = \int\limits_{4\pi} F(\varphi, \theta)\, d\Omega, \tag{I-77}$$

where $F(\varphi, \theta) = G(\varphi, \theta)/G_{max}$ is a function describing the antenna radiation pattern.

The following two particular cases are most frequently encountered in practice.

(1) The case of reception of radio emission from regions whose brightness temperature varies little within the limits of the antenna radiation pattern. In this case in expression (I-76) $T_e(\varphi, \theta)$ can be removed from under the integral sign; then $T_a = T_e$.

Thus, the source equivalent temperature referenced to the antenna is equal to the brightness temperature of the observed portion of the radio emitting region. This is usually the case with reception of the radio emission of the general background of the Galaxy, using highly-directive antennas.

(2) The case of the reception of radio emission of a source whose /81
angular dimensions are small in comparison with the width of the antenna radiation pattern ($\Omega_e \ll \Omega_a$). In this case we can consider that within the limits of the body angle of the source $G(\varphi, \theta) = G_{max} = \text{const}$; expression (I-76) is then reduced to the form

$$T_a = T_i \frac{\Omega_e}{\Omega_a}. \tag{I-78}$$

The source equivalent temperature referenced to the antenna can be expressed in this case in terms of the radio emission flux

$$T_a = \frac{PA}{2k}. \tag{I-79}$$

Expression (I-79) is more convenient for practical use, since for the determination of T_a we are required to know only one parameter of the source--the radio emission flux. This expression can be used for the calculation of T_a when the angular dimensions of the source are not known.

Let us turn to the consideration of the various forms of external noise. The galactic noise is the primary form of external noise. The intensity of the radio emission of the Galaxy depends on the coordinates and the wavelength. The most intense radiation emerges from the center of the Galaxy (right ascension α = 17 hr 50 min, declination $\delta = -28^{\circ}$), which is in the direction of the constellation Sagitta, and the least intense comes from the poles of the Galaxy.

Table I-12 presents the values of the intensity of the radio emission of the center of the Galaxy for various wavelengths, expressed in units of brightness temperature (ref. 49).

The distribution of radio brightness over the Galaxy also depends on the frequency. The representation of the brightness distribution of the cosmic radio emission over the sky for various frequencies is given by the isophotes--curves joining the different regions of the sky with equal brightness tem- /82
perature. Figure I-19 (a-i) presents the isophotes for the frequencies 64, 81, 100, 160, 250, 480, 600 and 910 Mcps (ref. 50) and the frequency 1,210 Mcps (ref. 51).

We note, first, that at the wavelengths of the centimetric and decimetric bands there is a pronounced concentration of radio emission toward the galactic equator and in the direction toward the center of the Galaxy. For example, at the 25 cm wavelength the extent of the region on the boundaries of which the intensity of the radio emission falls to half that at the center of the Galaxy amounts to about 10° in the direction of the plane of the Galaxy (along the Milky Way) and about 4° in the direction perpendicular to this plane. With increase of the wavelength, the maximum of the radio emission of the Galaxy broadens. At the λ = 62 cm wavelength the dimensions of the indicated region are already about 40° in the direction of the plane of the Galaxy and about 8° in the direction perpendicular to the plane of the Galaxy. At the wavelengths of the metric band the maximum of the radio emission becomes still more diffuse. For example, at the 3 m wavelength the longitudinal and lateral dimensions of the effective radiation region amount to 80 and 20°, respectively.

A second conclusion which can be made from an analysis of figure I-19 is that the contrast between the "dark" and "bright" portions of the sky diminishes with increase of the wavelength. For example, at the 3 m wavelength the radio brightness in the direction of the poles of the Galaxy is 10-12 times less than in the direction towards its center. However, at the 16.4 m wavelength the

TABLE I-12

f, Mcps	18.3	100	160	200	480	1,200	3,000
T_e, $^{\circ}$K	140,000	3,860	1,370	447	107	17	2.6

/83

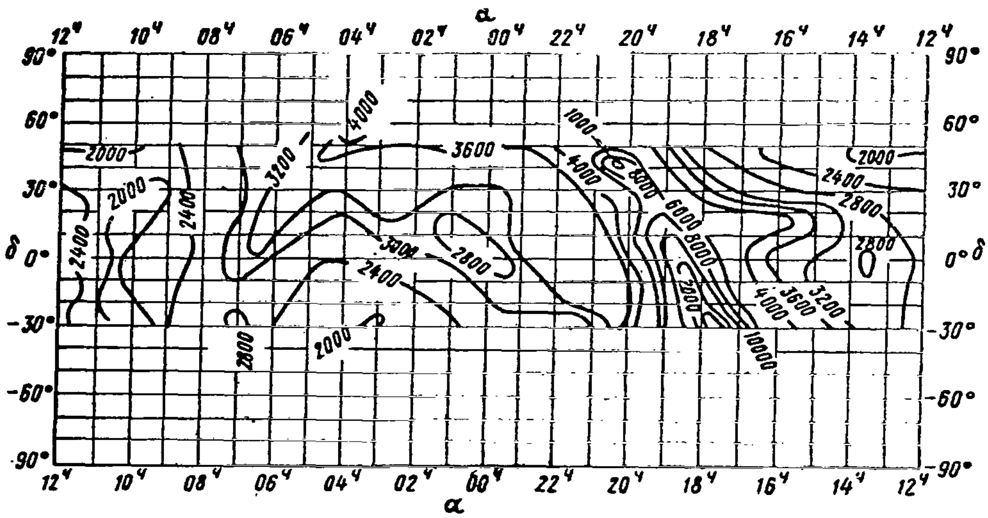


Figure I-19a. Isophotes of galactic radio emission at 64 Mcps. Numbers on isophotes indicate absolute temperature in $^{\circ}$K. (ч = hour).

/84

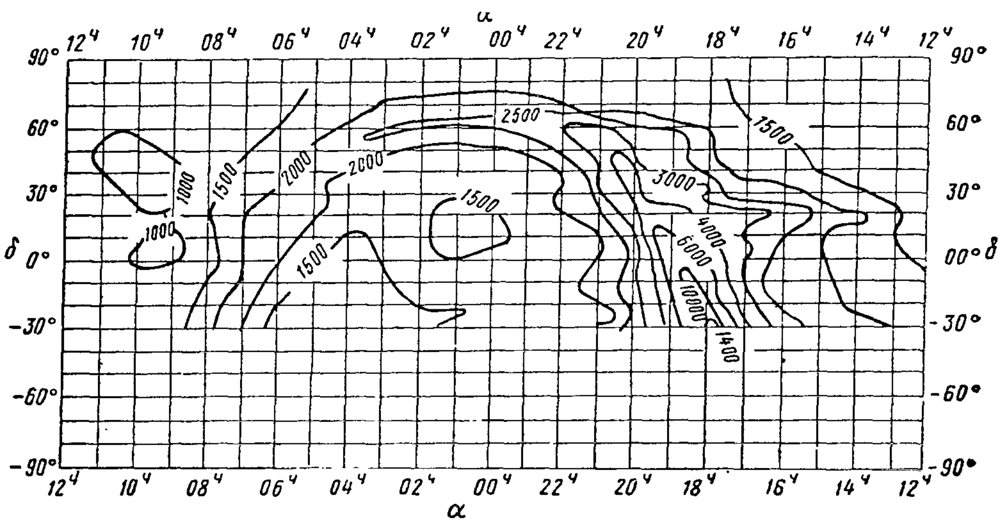


Figure I-19b. Isophotes of galactic radio emission at 81 Mcps. Numbers on isophotes indicate absolute temperature in $^{\circ}$K.

/85

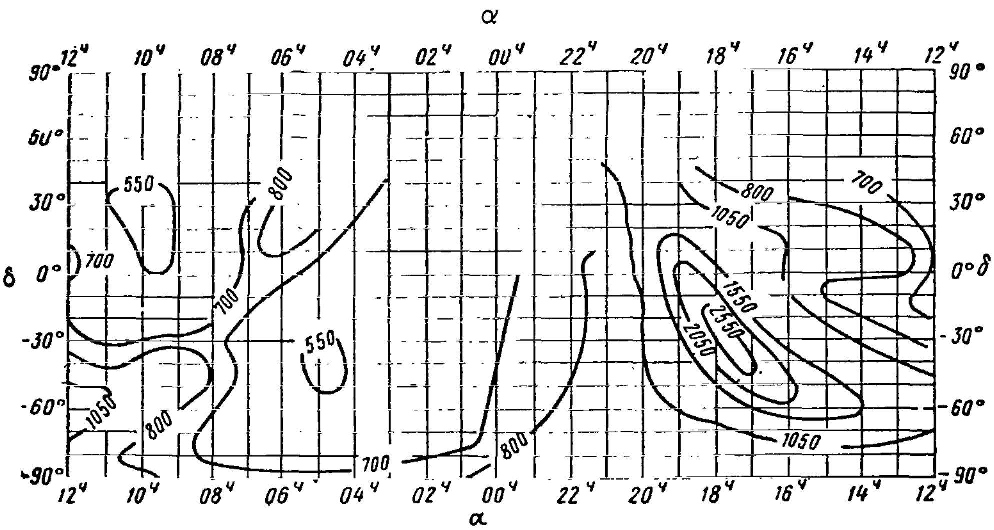


Figure I-19c. Isophotes of galactic radio emission at 100 Mcps. Numbers on isophotes indicate absolute temperature in °K.

/86

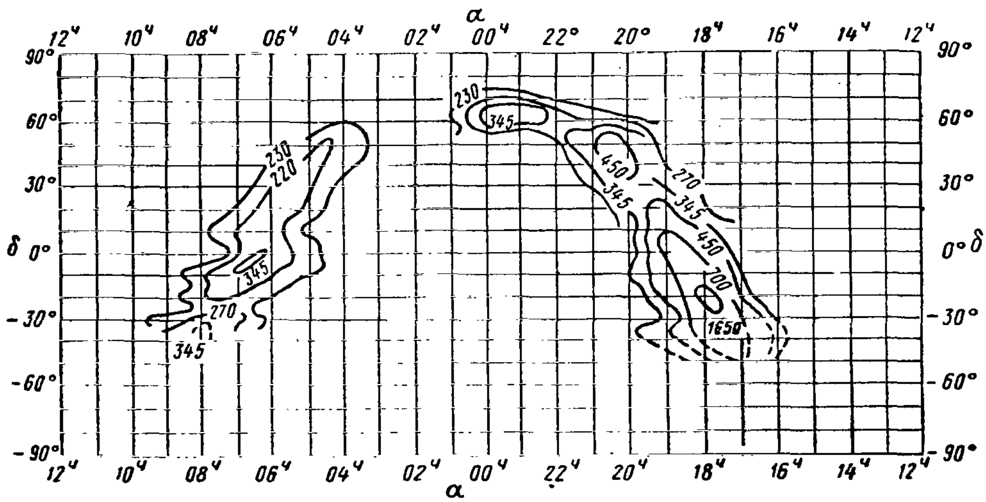


Figure I-19d. Isophotes of galactic radio emission at 160 Mcps. Numbers on isophotes indicate absolute temperature in °K.

/87

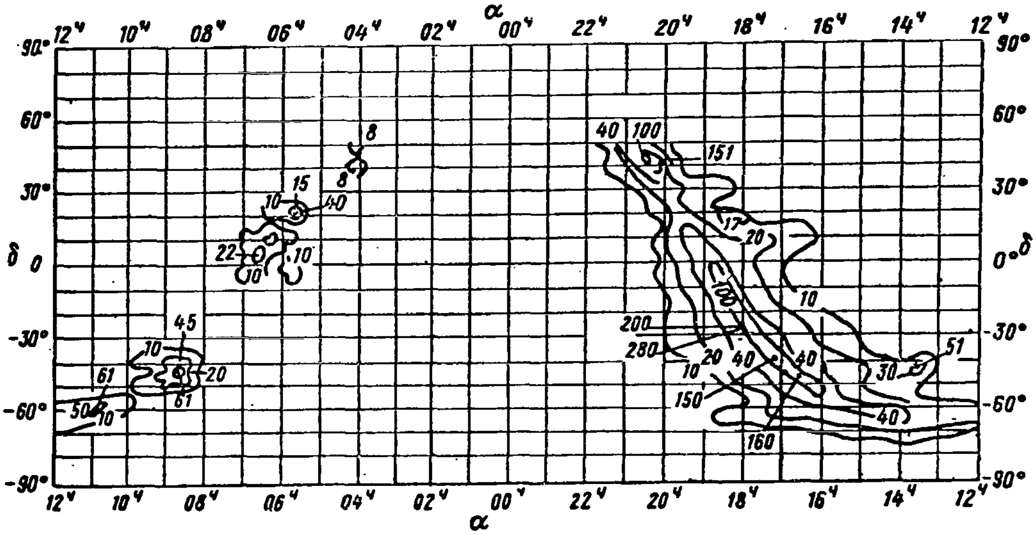


Figure I-19e. Isophotes of galactic radio emission at 250 Mcps. Numbers on isophotes indicate absolute temperature in °K.

/88

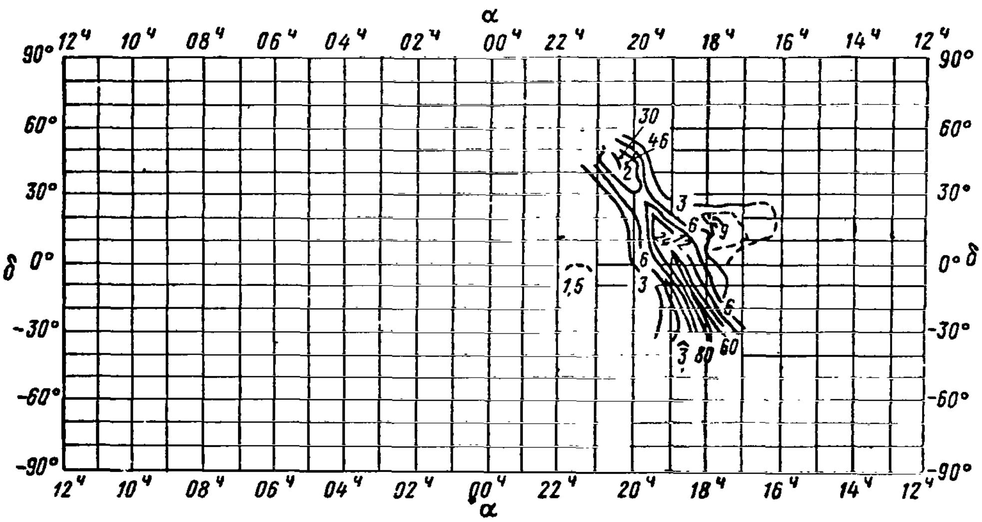


Figure I-19f. Isophotes of galactic radio emission at 480 Mcps. Numbers on isophotes indicate absolute temperature in °K.

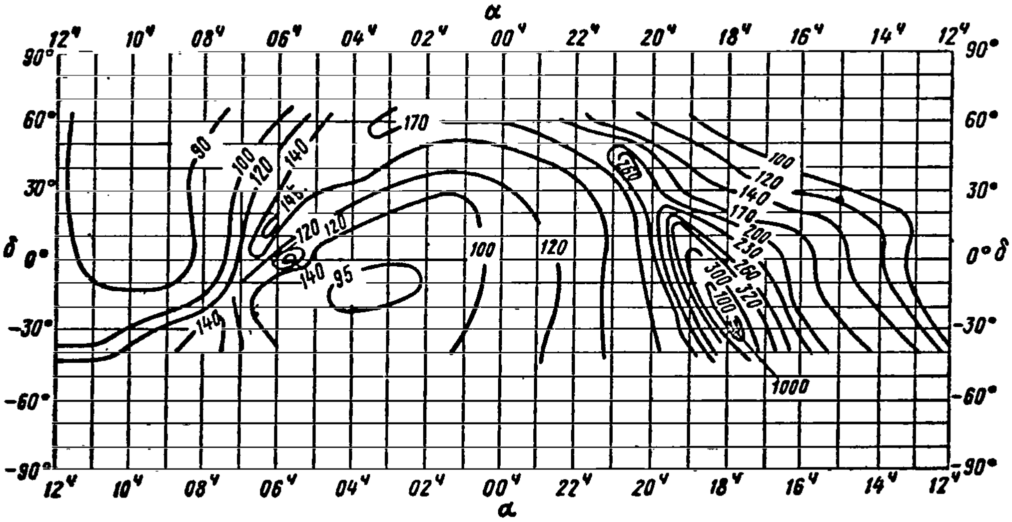


Figure I-19g. Isophotes of galactic radio emission at 600 Mcps. Numbers on isophotes indicate absolute temperature in °K.

/90

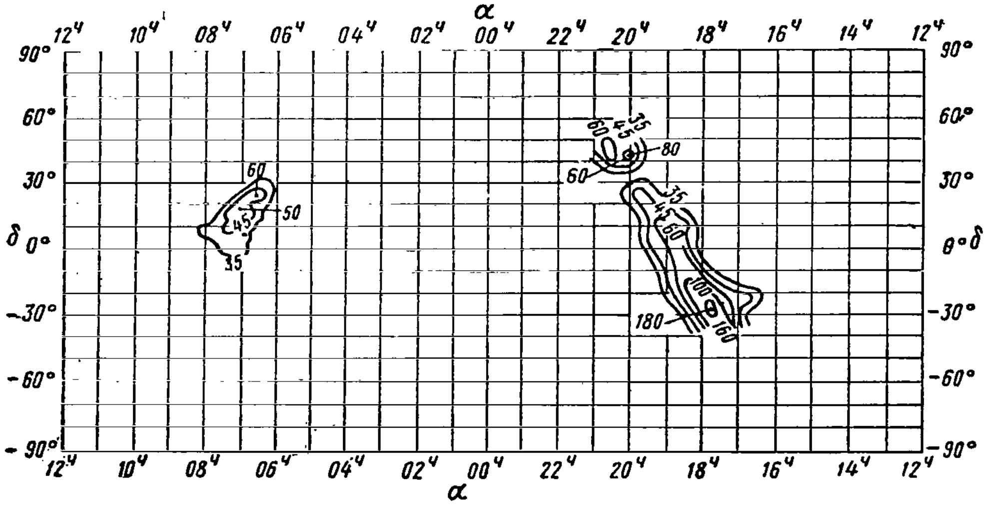


Figure I-19h. Isophotes of galactic radio emission at 910 Mcps. Numbers on isophotes indicate absolute temperature in °K.

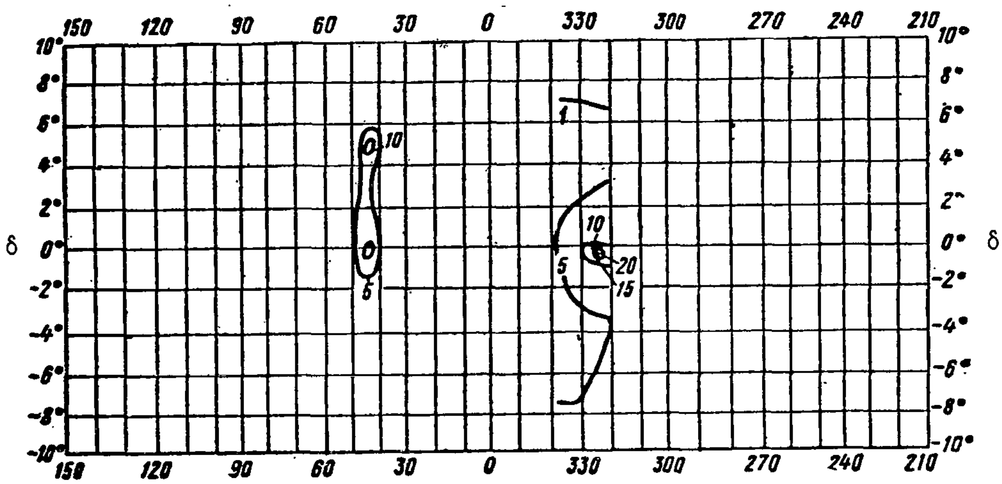


Figure I-19i. Effective antenna temperature in °K relative to "cold" portions of sky at 1,210 Mcps.

radio brightness in the direction toward the poles of the Galaxy is less than in the direction towards its center by a factor of 4-5.

A third conclusion concerns the spectral composition of the galactic radio emission. From figure I-19 we see a general reduction of the brightness temperature with increase of frequency. The spectral law for all regions of the sky has the form

$$T_e \sim f^{-\alpha}, \tag{I-80}$$

where α is approximately equal to 2.4 (definitely greater than 2).

In the calculations of the power requirements we shall assume that the most unfavorable situation is realized: the ground receiver antenna is directed toward the center of the Galaxy. If we use antennas with a narrow radiation pattern ($\Omega_a \ll \Omega_e$), then even at the high frequencies with /92 pointing toward the center of the Galaxy $T_a = T_e$. The curve of the variation of the brightness temperature of the center of the Galaxy as a function of wavelength can be described with an adequate degree of accuracy by equation

$$T_e = \lambda^{2.4} \cdot 469^\circ \mathrm{K}, \tag{I-81}$$

where λ is measured in meters.

Another form of cosmic noise is the noise of the discrete radio sources. At the present time a large number (about 2,000) discrete sources of radio emission are known (ref. 49). The flux of the radiation from these sources, just as from the Galaxy, increases with an increase of the wavelength.

Table I-13 presents the coordinates and intensity of radio emission of eight of the most powerful discrete sources observable from the territory of the USSR.

The dimensions of the discrete sources are small, and they can be considered to be point sources, since only about 40 of the discrete sources have angular dimensions of more than 20'. In view of the small angular dimensions of the most powerful discrete sources (of the order of a few angular minutes) the probability of happening on them in the ground communication antenna pattern is low, if this antenna is large (narrow radiation pattern). If the directivity coefficient of the ground antenna is not large (broad radiation pattern), the occurrence of one or even several discrete sources in the field of such an antenna presents no danger, since their contribution to the equivalent noise temperature at the receiver input is small. For example, with operation using an antenna having an effective area $A = 1\ m^2$ at the wavelength $\lambda = 20$ cm, the equivalent noise temperature of the most powerful discrete source Cygnus A does not exceed $0.9^{o}K$, which is considerably less than the temperature of the center of the Galaxy (about $17^{o}K$). Therefore, in the determination of the equivalent noise temperature at the receiver input we shall not take account of the noise of the discrete radio emission sources.

A powerful source of cosmic radio noise is the Sun. We differentiate the radio emission of the quiet Sun, observed in the absence of spots on the surface of the Sun, and the radio emission of the noisy Sun, observed with the presence of such formations. /94

The intensity of the radio emission of the quiet Sun is quite constant from day to day and in the ranges of the metric and millimetric waves is the same for years of maxima and minima of the solar activity. In the range of the centimetric and particularly the decimetric waves the intensity of the radio emission of the quiet Sun varies in accordance with the 11-year cycle of solar activity. The average data on the intensity of the radio emission of the Sun in the range from 8 mm-10 m are presented in table I-14 (ref. 49). The numbers in the numerator correspond to the years of a maximum and those in the denominator to years of a minimum of the solar activity.

The radio emission of the noisy Sun is characterized by a general increase of the intensity of the radiation and by strong variations of this radiation. This increase depends on the wavelength, and on the average amounts to several percent in the range of the millimeter waves (MMW); several tens of percent in the range of the centimeter waves (CMW); factors of 1.5-3 in the range of the decimeter waves (DMW); and factors of tens and hundreds of times in the range of the meter waves (MW). The duration of the disturbances varies from several minutes in the MMW range to several days in the MW band.

TABLE I-13

Source	Coordinates		Radio emission flux, $P \cdot 10^{+24}$, W/m^2 cps						
	Right ascension α	Declination δ	λ = 3.2 cm	λ = 10 cm	λ = 20 cm	λ = 50 cm	λ = 100 cm	λ = 300 cm	λ = 1,000 cm
Cassiopeia	23 h 21 min	+ 58° 30'	5.9	15	25	40	60	150	600
Cygnus A	19 h 57 min	+ 40° 35'	-	8	12	20	35	110	400
Taurus	05 h 41 min	+ 22° 04'	7.3	8	10	13	16	18	18
Virgo	12 h 28 min	+ 12° 44'	-	1.8	2.3	3	5	12	-
Centaur A	13 h 22 min	- 42° 46'	-	2.2	2.8	4.5	7	18	-
Orion M42	5 h 33 min	- 5° 37'	2.7	4.5	4.5	-	-	-	-
Nebula Omega M17	18 h 17 min	- 16°	7.5	7	8	-	-	-	-
Nebula M20	17 h 59 min	- 23°	-	1	4	-	-	-	-

TABLE I-14

Wavelength	8 mm	3 cm	10 cm	25 cm	50 cm	1.5 m	3 m	10 m
Radiation flux $P \cdot 10^{21}$, W/m^2 cps	200	$\frac{32}{27}$	$\frac{13}{6.5}$	$\frac{7}{3.5}$	$\frac{5}{2.5}$	0.85	0.23	0.035

In addition to this relatively slight increase of intensity, we infrequently (several times a year) observe particularly strong bursts of radio emission, during which the intensity increases by 20-30 percent in the MMW band, by tens of times in the CMW band, by hundreds of times in the DMW range, and by /95 tens and hundreds of thousands of times in the MW band. The duration of these bursts is from several minutes to an hour.

The relation between the periods during which the radio emission of the Sun has a quiet or noisy nature depends on the phase of the 11-year cycle of the solar activity. In the years of maximum solar activity the radio emission of the Sun has a disturbed nature up to 30-50 percent of the time. In years of minimum solar activity the number of days in which the radio emission of the Sun has a disturbed nature amounts to less than 10 percent. The last maximum of the solar activity occurred in 1958; 1964 is a year of minimal solar activity.

Let us compare the intensity of the solar noise with the galactic noise. In a year of minimal solar activity the radio emission flux of the Sun at the 25 cm wavelength is $3.5 \cdot 10^{-21}$ W/Mcps. With reception on an antenna with an effective area $A_1 = 1$ m^2, the antenna temperature resulting from the solar emission is $T_{a_1} = 125^\circ$K; if the area $A_2 = 10$ m^2, then $T_{a_2} = 1{,}250^\circ$K; if the area $A_3 = 100$ m^2, then $T_{a_3} = 12{,}500^\circ$K. These values considerably exceed the antenna temperature resulting from the galactic noise, even with the antenna directed at the center of the Galaxy ($T_{a\ gal} = 17^\circ$K). Still higher values are obtained for years of maximal solar activity and the disturbed periods.

With the use of large receiving antennas, the appearance of the Sun in the field of the antenna leads to complete "blinding" of the receiving station. If a molecular amplifier is installed at the ground receiver input, it may be damaged. Therefore this possibility must be taken into account, and in the compiling of the program for the antenna travel (if satellite tracking by the antenna is programmed) consideration must be given to the position of the Sun, although because of the relatively small angular dimension of the Sun (0.5°) the probability of it falling in the antenna pattern of a large ground antenna (with narrow beam width) is small (ref. 125). In certain cases, for example with a maser at the receiver input, it may be advisable to switch the receiver off during the time of passage of the Sun through the field. /96

Earth and the planets are also sources of radio noise. The surface of Earth radiates radio noise with an equivalent temperature of about 300°K. This must be taken into account in the design of the Earth-satellite radio link. For the downlink the receiver is located on Earth and reception of Earth's radiation can occur only by the side lobes of the field pattern, and only if the main lobe is quite narrow and the antenna elevation angle is other than 0°. Usually the level of the side lobes does not exceed 1 percent (in power) of the level of the main lobe (ref. 52). Consequently, the antenna noise temperature due to the thermal radio emission of Earth will not exceed 30°K.

The surface temperature of the planets is comparable with the temperature of Earth. However, due to the small angular dimensions, their contribution to the equivalent noise temperature at the receiver input is negligibly small and can be ignored in the design of the radio link. In addition, the probability of the appearance of the planets in the ground receiving antenna field is also small. The nonthermal emission of the planets has a sporadic nature. The fluxes of nonthermal radio emission of the planets are small and can be ignored in the power requirement computations of the radio links being analyzed. In addition to the thermal radio emission of Earth, we must also take account of the noise of Earth's atmosphere. The atmospheric noise at the high frequencies is due to absorption of the Sun's radiant energy (and its reradiation) by the oxygen and water vapor. The intensity of this form of noise depends to a considerable degree on the elevation angle δ of the receiving antenna above the horizon and is maximal with $\delta = 0^{\circ}$, since in this case the radiating volume includes a greater mass of the atmospheric components mentioned above. An idea of the frequency and angular distribution of this form of noise is given by the curves of figure I-20 ($\delta = 0, 5, 10, 30, 90^{\circ}$), where the effective radiation temperature of the troposphere is shown as a function of the elevation angle and the frequency, as computed by Hogg and Mumford (ref. 17). Here we show, for comparison, the curves of the variation of the effective temperature of the center and the poles of the Galaxy (max and min) as a function of frequency. We can see that the atmospheric noise shows up at frequencies of 1,000-2,000 Mcps and higher. In this range it predominates over the galactic noise and is practically the only form of external noise which must be accounted for in the selection of the operating frequency and in the power requirement calculations. /97

We see the strong variation of the intensity of the atmospheric noise with elevation angle in the region of small values of this angle. For example, with increase of δ from 0-5° the equivalent noise temperature diminishes by about a factor of 2-3 in the frequency range of 500-10,000 Mcps. With further increase of δ the reduction of the intensity of the atmospheric noise proceeds more gradually.

It was shown previously that the time of simultaneous visibility of a satellite from two or more points of Earth's surface is one of the most important parameters of the communication system using the AES, and depends to a considerable degree on the minimal elevation angle of the ground antenna δ_{min}.

From the point of view of improving the efficiency of the communication system, it is desirable to reduce δ_{min}. The selection of the optimal value of δ_{min} depends on a large number of parameters. The strong variation of the intensity of the atmospheric noise with δ for $\delta < 5^{o}$ and its more gradual decrease with $\delta > 5^{o}$ is one of the arguments in favor of the selection of the minimal angle elevation of the ground antenna in the vicinity of $\delta_{min} \approx 5^{o}$. /98

We mentioned above that with operation of an antenna with low elevation angles it is possible to encounter noise at the receiver input caused by the radiation of Earth as a blackbody with a temperature of about 300°K.

However, the operation of the antenna with small elevation angles is dangerous from points of view other than that of the appearance of this noise. There is a more serious danger--the possibility of the occurrence of strong fading of the signal as a result of the arrival of one or several beams reflected from the surface of Earth, in addition to the primary beam. As the result of the motion of the satellite there is a change of the phase difference of the direct and reflected rays; because of the unevenness of the surface of Earth there is a change of the amplitude of the reflected ray since there is a change of the angle of incidence of the reflected ray, and as a result of the finite width of the field pattern there will be sharp variations of the intensity of the received and reflected signal.

Thus, the amplitude of the total signal will fluctuate. The intensity of the fluctuations is determined by the width of the receiving antenna pattern (intensity increases with greater pattern width) and by the magnitude of the reflection coefficient of the radio wave from Earth's surface. The magnitude of the reflection coefficient of the UHF band wavelength from Earth's surface

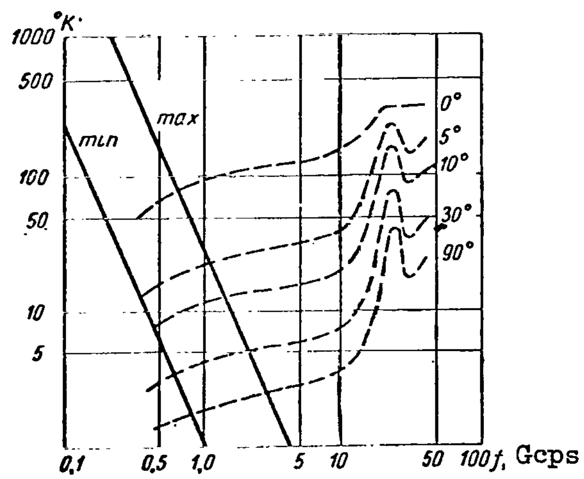


Figure I-20. Atmospheric noise in °K with various elevation angles as function of frequency.

depends on the type of reflecting surface and the form of polarization of the radio wave.

In the case considered of grazing propagation, there are particularly favorable conditions for the reflection of the wave, since the reflecting region (first Fresnel zone) in practice is quite smooth and uniform because of its smallness. In this case larger values of the reflection coefficient are obtained for the horizontally polarized waves. Reference 53 presents experimental data on the variation of the effective reflection coefficient ρ_{eff} as a function of the angle ψ with the reflecting surface for vertically and horizontally polarized waves at the $\lambda = 9$ cm and 26.5 cm wavelengths for reflection from land (fig. I-21, curves a and b). With small angles ψ we have $\rho_{eff} \approx 0.9$ for both polarizations. If the width of the antenna pattern is sufficiently large, then in the worst case the amplitude of the overall wave (with phase-opposition /99 addition of the beams) is equal approximately to 0.1 A_0, where A_0 is the amplitude of the direct wave, i.e., there is a power attenuation of the signal by a factor of 100.

Let us estimate the greatest possible width of the main lobe of the receiving antenna pattern, assuming the height of the phase center above the surface of Earth to be Δh. Figure I-22 shows that the most favorable conditions for the reception of the reflected beam with a given width of the receiving antenna pattern are obtained for the rays incident on the line of intersection of the boundary AD of the zone of visibility from the phase center B (horizon line) with the projection of the antenna pattern on Earth.

Note. This case is unrealizable with an ideally smooth surface of Earth; however the actual surface of Earth, for the frequency range of interest to us, is not an ideally smooth reflector--it has roughnesses and slopes. For this reason, on the surface of Earth ahead of the antenna there is always an area the size of the first Fresnel zone (of the order of 1 m), having an angle such that a radio beam incident on it is reflected in the direction of the antenna. /100

It is evident that the maximal width of the pattern at the half-power points θ_{max} is determined from the relation

$$\frac{\theta_{max}}{2} \leq \alpha + \delta. \tag{I-82}$$

From the triangle AOB we find that

$$\begin{aligned} \alpha &= 90^\circ - \angle ABO, \\ \angle ABO &= 90^\circ - \angle AOB; \\ \angle AOB &= \arcsin\left(\frac{S}{R_0 + \Delta h}\right), \end{aligned} \tag{I-83}$$

where S is the length of the segment AB, equal to $\sqrt{\Delta h R_0}$.

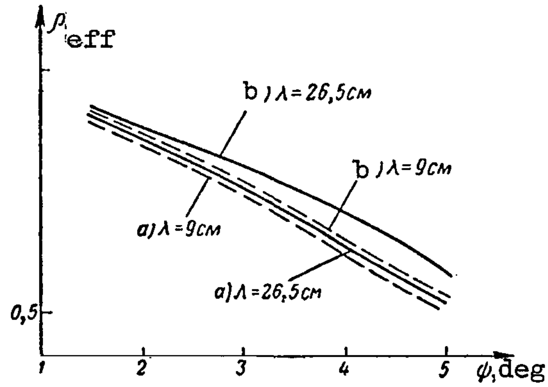


Figure I-21. Variation of effective reflection coefficient ρ_{eff} as function of encounter angle ψ at various wavelengths.

a, vertical polarization (reflection from land); b, horizontal polarization (reflection from land).

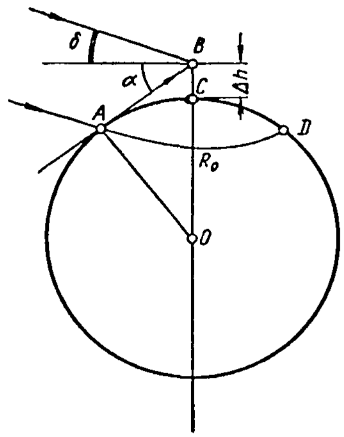


Figure I-22. Paths of direct and reflected (from irregularities of Earth's surface) rays.

Even in the case $\Delta h = 50$ m the value of α does not exceed 11'. We have limited ourselves to the value $\delta_{min} = 5^{\circ}$. Therefore, the maximal value of the width of the receiving antenna pattern at the half-power points is equal to $\theta_{max} = 10^{\circ}$. With a given diameter D_2 of the ground-based antenna, accounting for the multipath effect leads to a limitation of the operating wavelength

$$\lambda_{op} \leq \lambda_{crit} (D_2, \delta_{min}).$$

Since the width of the pattern at the half-power points θ is determined from equation (ref. 54)

$$\theta^{o} \approx \frac{70\lambda}{D_2}, \tag{I-84}$$

the critical wavelength is determined from condition

$$\lambda_{crit} = 0.143 \cdot D_2 \tag{I-85}$$

If, however, the condition $\lambda_{op} < \lambda_{crit}$ is satisfied, then condition /101 $1/2\ \theta - \delta_{min} = 0$ is automatically satisfied, and Earth noise for practical purposes does not affect the receiver input. Because of the indefinite nature of the value of the fluctuating signal amplitude resulting from multipath transmission, we shall not take it into account in the calculations of the power required, but in the computations of the bandwidth of the quasioptimal wavelengths we shall note the boundary of the critical range where fading can occur.

The consideration of the various sources of external noise makes it possible to draw the following conclusions.

1. In the determination of the optimal operating wavelength and the power requirements we need consider only the galactic and atmospheric noises.

2. In setting up the program for the movement of large ground-based antennas (with high directivity coefficients) it is necessary to take account of the position of the Sun, and in the case of the use of a maser at the input of the ground receiver it will be necessary to disconnect the receiver from the antenna during the time of passage of the Sun through the field (ref. 125).

Thus, the value of T_{eff} in the communication equation is written in the following way

$$T_{eff} = T_0 + \Gamma\, T_{a_{gal}} + T_{a_{atm}},$$

where T_0 is the equivalent internal noise temperature of the receiver (values are assumed of 50, 100, 300, 600, 1,500°K);

$T_{a_{gal}}$ is the equivalent temperature of the galactic noise (taken for the center of the Galaxy);

$T_{a_{atm}}$ is the equivalent temperature of the atmospheric noise ($T_{a_{atm}}$ taken for $\delta = 5^{o}$).

Frequency Shift Due to Doppler Effect and Various Instabilities. The RF passband width, directly proportional to the required radiation power P_t, is equal to the active width of the spectrum of the signal F_0 plus the frequency shift due to the Doppler effect Δf_D and to the instability of the transmitter frequency and the receiver heterodyne Δf_{des}. Thus, the value of $\Delta f_D + \Delta f_{des}$ is important for the calculations of the power requirements. /102

The magnitude of the frequency shift resulting from the instability of the transmitter and the receiver heterodyne is determined by the radio link frequency instability ν_{des}

$$\Delta f_{des} = \nu_{des} f_0, \tag{I-86}$$

where f_0 is the carrier frequency.

Let us find the expression for the Doppler shift of the frequency. As is known, in the case of the motion of a source of vibrations toward the observer with the velocity v_1, the frequency of the vibrations received at the observer's location will be

$$f_1 = \frac{f_0}{1 - \frac{v_1}{c}}, \tag{I-87}$$

where f_0 is the frequency of the vibrations emitted by the source;

c is the velocity of propagation of these oscillations in the medium (in the present case the velocity of light).

For satellite velocities $v_1 \leq 10$ km/sec, the condition $v/c \ll 1$ is satisfied. In this case

$$\frac{1}{1 - \frac{v_1}{c}} \approx 1 + \frac{v_1}{c}$$

and

$$\Delta f_d \approx f_0 \frac{v_1}{c}. \tag{I-88}$$

The Doppler frequency shift is nonzero, if the projection of the satellite velocity on the direction of the radius-vector "observer-satellite" (radial component of the velocity) $v_{D_i} = -v_1$ is nonzero.

It is evident that

$$V_{d_i} = \frac{dR_i}{dt},$$

where R_i is the distance between the satellite and the point of observation.

For the case of the circular polar orbit and an observer located at the North Pole, the following expression is obtained for v_1 /103

$$v_1 = \frac{(R_0 + h) R_0 \omega_c \sin \omega_c t}{\sqrt{(R_0 + h)^2 + R_0^2 - 2(R_0 + h) R_0 \cos \omega_c t}}, \tag{I-89}$$

where R_0 is the radius of the Earth;

h is the satellite flight altitude above Earth;

$\omega_c = 2\pi/T_c$ where T_c is the period of revolution of the satellite in orbit.

In the majority of the cases we are interested in the maximal shift of the frequency as the result of the Doppler effect. It is evident that the Doppler frequency shift is maximal at the moment of radiorise and radioset of the satellite relative to the ground observer. The condition of radiorise (radioset) of the satellite for the ground observer can be written most simply for the case $\delta = 0^{\circ}$.

In this case the times of "rise" and "set" t_1 and t_2 are found from equation

$$t_{1,2} = \mp \frac{\arccos \frac{R_0}{R_0 + h}}{2\pi} T_c \tag{I-90}$$

(as $t = 0$ we take the time of passage of the satellite over the North Pole). Substituting $t_{1,2}$ in the equation for v_1, we find $v_1 = v_{1max}$. For h = 2,000, 4,800 km/sec we find, respectively,

$$v_{1max} = 5.25;\ 3.52 \text{ km/sec}.$$

In the case of an elliptic orbit inclined at an angle of 63°, the equation for the determination of v_1 is found to be quite cumbersome, although the difference from v_1 for the case of the polar elliptic orbit will be slight. Therefore, we shall determine v_1 for the polar elliptic orbit, taking the same parameters as for the elliptic orbit. We shall carry out the calculations for the case of the observer at the North Pole. Here we assume that the satellite is used for communications purposes only in the course of that half of /104 the period when it is near the apogee (fig. I-23). In this case the Doppler shift will be maximal at the time the satellite is at point P_2. Let us find v_1 for this point.

The position of the satellite in orbit at any instant of time t is given by its radius-vector R_c, drawn from the center of the Earth O, and by the angle f with vertex at O between R_c and the line of apsides (ref. 20). The angle f is termed the true anomaly. The value which the true anomaly would have if the satellite moved around the center of Earth with a constant angular velocity n, equal to $2\pi/T_c$, where T_c is the period of revolution of the satellite, is termed the mean anomaly g.

The mean anomaly g, equal to nt (where t is reckoned from the moment the satellite passes through the perigee), and the radius-vector R_c are expressed as follows in terms of the true anomaly (I-8), (I-9)

$$R_c = \frac{a(1-e^2)}{1+e\cos f},$$

$$g = \arc\cos\left[\frac{1}{e} - \frac{1-e^2}{e(1+e\cos f)}\right] - e\sqrt{1-\left[\frac{1}{e}-\frac{1-e^2}{e(1+e\cos f)}\right]^2}$$

The distance between the satellite and the observer at the Pole is /105 found from equation

$$R = \sqrt{R_0^2 + R_c^2 - 2R_cR_0\sin(f-90^\circ)}. \tag{I-91}$$

Differentiating with respect to t, we find

$$v_1 = \frac{R_0\dot{R}_c\sin(f-90^\circ)}{R} + \frac{R_cR_0\cdot\dot{f}\cos(f-90^\circ)}{R} - \frac{R_c\dot{R}_c}{R}, \tag{I-92}$$

where

$$\dot{R}_c = \frac{2\pi a e \sin f}{\sqrt{1-e^2}\,T_c};$$

$$\dot{f} = \frac{\pi a(1 + e\cos f)}{R_c\sqrt{1-e^2}}\,\frac{2\pi}{T_c}.$$

In the example considered on page 14 the parameters of the inclined elliptic satellite orbit are the following: apogee altitude h_a = 20,000 km, perigee altitude h_p = 500 km. Hence we can find the semimajor axis, the eccentricity and the period: a = 16,650 km, e = 0.585, T_c = 5 hr 50 min. The point P_2 of the satellite orbit corresponds to g = 90°. Thus we find that f = 150°, R_c = 20,700 km. Substituting the values of all known quantities in the equation for v_1, we find that v_1 = 2.42 km/sec.

It is evident that for both active and passive relaying the Doppler shift is maximal for the case of the communication of two nearby corresponding stations. In this case it is necessary to expand the ground-based receiver passband by the amount

$$\Delta f_d = \frac{4v_1}{c} f_0 = v_d f_0, \tag{I-93}$$

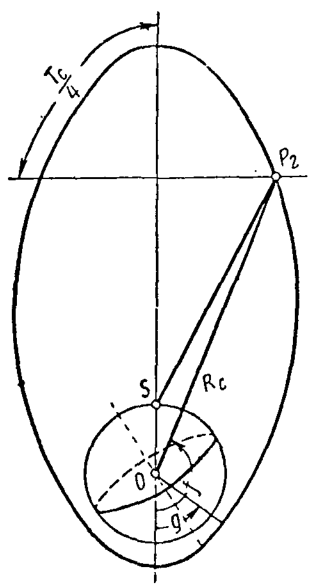


Figure I-23. Geometry of problem of determination of Doppler frequency shift.

where ν_D is the coefficient characterizing the required expansion of the ground receiver passband because of the Doppler shift, and it is considered that at the moments of radiorise and radioset the Doppler frequency shift has opposite signs.

For the circular orbits with altitudes of h = 2,000 km and 4,800 km, the coefficient ν_D takes the values $7 \cdot 10^{-5}$ and $4.68 \cdot 10^{-5}$, respectively. For the elliptic orbit with h_a = 20,000 km, $\nu_D = 3.24 \cdot 10^{-5}$.

Let us assume that the frequency stability is 10^{-4} and that the Doppler frequency shift is not compensated. In this case the coefficient ν_Σ, characterizing the overall frequency shift of the radio link, take the values $\nu_{\Sigma_1} = 1.7 \cdot 10^{-4}$; $\nu_{\Sigma_2} = 1.47 \cdot 10^{4}$; $\nu_{\Sigma_3} = 1.32 \cdot 10^{-4}$; $\nu_{\Sigma_4} = 1 \cdot 10^{-4}$, respectively for the circular orbits with h = 2,000 km and 4,800 km (refs. 17, 18), for the elliptic orbit with h_a = 20,000 km (ref. 15), and for the stationary orbit (for it the Doppler shift is equal to 0).

Thus, the coefficient F appearing in the communication equation is written as

$$F = F_0 + \nu_\Sigma f_0,$$

where F_0 is the RF signal spectrum width (determined by the form of information and method of modulation);

ν_Σ is the coefficient characterizing the overall shift of the radio link frequency (taken for the case of the maximal Doppler shift for each orbit).

Margin Coefficient Γ. The margin coefficient Γ characterizes the effect of the medium (Earth's atmosphere) on the radio waves propagating through it. Here two regions of the terrestrial atmosphere have the greatest effect on the electromagnetic waves--the ionosphere and the troposphere (ref. 124).

Effect of Ionosphere on Radio Wave Propagation. The ionizing regions of the terrestrial atmosphere, located more than 60 km above the surface of Earth, divided into several layers (D, E, F_1 and F_2) and combined by the single term "ionosphere," have a significant effect on the propagation of radio waves through them.

The higher the electron concentration in the ionosphere and the longer the operating wavelength, the more it is affected by the ionosphere. If the frequency of the radio wave is below the critical value for one of the ionospheric layers, the wave cannot pass through the ionosphere: it is reflected. The electron concentration, which exists at a height of h = 60-90 km (D layer, $N_e \approx 10^2$ cm^{-3}), is sufficient to reflect the long wavelengths transmitted from Earth. The medium wavelengths are reflected by the E layer, located at a height of 100-130 km, where the electron concentration is greater than ($N_e \approx 10^5$ cm^{-3}). The still shorter wavelengths, after passing through the D and E layers, are reflected by the F layer with a maximal electron concentration of $N_e \approx 10^6$ cm^{-3} at an altitude h = 200-400 km. /107

The critical frequencies of the F layer lie in the range of 5-15 Mcps, depending on the season, time of day and solar activity cycle (ref. 5). The reflection frequency with inclined incidence on this layer is

$$f = \frac{f_{cr}}{\cos\psi}, \tag{I-94}$$

where ψ is the angle of incidence. With operation using an antenna having an elevation angle $\delta = 0^\circ$, radio waves with a frequency of f = 20-50 Mcps are reflected from the ionosphere. As a result of the presence in the atmosphere of large-scale formations with high electron concentration gradients at altitudes of 130-150 km (E_S layer), the higher frequencies right up to 100 Mcps may also be reflected.

Thus, communication with space vehicles located beyond the limits of the atmosphere must be maintained at frequencies higher than 100 Mcps.

However, at such high frequencies the effect of the ionosphere, as before, is large and manifests itself in the absorption of the radio waves on passage through the ionospheric layers, scattering on the ionospheric inhomogeneities and change of direction of the polarization plane on passage through the atmosphere (Faraday effect). Let us consider each of these effects separately.

Absorption of Radio Waves in the Ionosphere. In the approximation of geometric optics the electromagnetic wave in a medium at level z can be written in the form

$$\overline{E} = E_0 \exp\left[-i \int_{z_0}^{z} \frac{\omega}{c} (n_{1,2} - i\chi_{1,2})\, dz\right], \tag{I-95}$$

where E_0 is the wave amplitude at the z level;

ω is the wave cyclic frequency;

c is the speed of light; /108

$\chi_{1,2}$ is the absorption coefficient (subscript 2 relates to the ordinary wave, subscript 1 to the extraordinary wave);

$n_{1,2}$ is the wave refraction index.

The complex index of refraction $(n_{1,2}-i\chi_{1,2})$ is expressed in terms of the parameters of the medium (case of normal incidence of the wave on the ionospheric layer) (ref. 5)

$$(n - i\chi_{1,2})^2 = 1 - \frac{2v \times (1-v-is)}{2(1-is)(1-v-is) - u\sin^2\alpha \pm \sqrt{u^2\sin^4\alpha + 4u(1-v-is)^2\cos^2\alpha}}, \qquad \text{(I-96)}$$

where

$$u = \left(\frac{f_H}{f}\right)^2;$$

$$v = \left(\frac{f_0}{f}\right)^2;$$

$$s = \frac{\nu_{eff}}{2\pi f};$$

f_H is the electron gyromagnetic frequency in the ionosphere, equal to $|e| H_0/2\pi mc$;

e is the electron charge;

H_0 is the intensity of the Earth's magnetic field at altitude z;

m is the electron mass;

f_0 is the ionospheric plasma frequency, equal to $\sqrt{e^2 N_e/\pi m}$;

N_e is the electron concentration at altitude z;

f is the frequency of the wave incident on the ionospheric layer;

ν_{eff} is the effective electron collision frequency at altitude z;

α is the angle between $\overline{H}_0$ and the z axis.

The X, Y, Z coordinate system is chosen so that the z axis is normal to the Earth's surface and the magnetic field vector $\overline{H}_0$ lies in the y, z plane (fig. I-24). The equation for n-iχ takes a relatively simple form in the case of longitudinal propagation (α = 0)

$$(n - i\chi)^2_{1,2} = 1 - \frac{v}{1 - is \pm \sqrt{u}}. \qquad \text{(I-96')}$$

We can show that in the case of interest to us of high frequencies (f >> f_H) the condition of quasilongitudinal propagation is satisfied in a wide range of angles α. Actually, in the case f >> f_H /109

$$u \gg u^2, \ \sqrt{u} \gg u$$

and approximately

$$(n - i\chi)^2_{1,2} \approx 1 - \frac{v}{1 - is \pm \sqrt{u} \cos\alpha}. \qquad \text{(I-96")}$$

For the range f = 100-1,000 Mcps the condition of quasiaxial propagation is satisfied with an accuracy to 5 percent over a wide interval of angles from 0 to 82-89°55', i.e., over practically the entire range of variation of the angle α.

It is evident that equation (I-96') can be rewritten in the form (we separate the real and imaginary parts)

$$\left.\begin{aligned} n^2_{1,2} - \chi^2_{1,2} &= 1 - \frac{v(1 \pm \sqrt{u}\cos\alpha)}{(1 \pm \sqrt{u}\cos\alpha)^2 + s^2}, \\ 2n_{1,2}\chi_{1,2} &= \frac{vs}{(1 \pm \sqrt{u}\cos\alpha)^2 + s^2}. \end{aligned}\right\} \qquad \text{(I-97)}$$

Since at the high frequencies $\sqrt{u}\cos\alpha \ll 1$, f >> f_0 and therefore /110 $n_{1,2} \approx 1$, $\chi_{1,2} \ll n_{1,2}$, then

$$\left.\begin{aligned} n_{1,2} &= 1 - \frac{v(1 \pm \sqrt{u}\cos\alpha)}{(1 \pm \sqrt{u}\cos\alpha)^2 + s^2}, \\ \chi_{1,2} &= \frac{1}{2}\,\frac{vs}{(1 \pm \sqrt{u}\cos\alpha)^2 + s^2}. \end{aligned}\right\} \qquad \text{(I-98)}$$

Since $\sqrt{u}\cos\alpha \ll 1$, $s \ll 1$ we obtain finally

$$\chi_{1,2} \approx \frac{1}{2} vs = 0{,}65 \cdot 10^7 \frac{N_e \nu_{eff}}{f^2}, \qquad \text{(I-99)}$$

where N_e is reckoned in electrons per cm^3, ν_{eff}, in sec^{-1} and f in cps. With passage through the entire thickness of the ionosphere, the integral absorption (power) coefficient is equal to

$$\mu = 2\int_{z_0}^{z} \frac{\omega}{c} \chi_{1,2} dz = 3 \cdot 10^{-3} f^2 \int_{z_0}^{z} N_e \nu_{eff} dz. \qquad \text{(I-100)}$$

Thus, the overall losses of the wave are determined by the frequency of the vibrations of the incident wave (at the high frequencies, inversely proportional to the square of the frequency), the electron concentration N_e and the average number of electron collisions ν_{eff}. The values of N_e and ν_{eff} at the various altitudes from z = 60 km to z = 1,000 km are presented in table I-15.

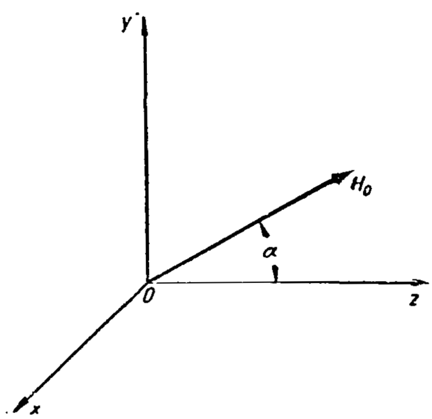


Figure I-24. Coordinate system for calculation of integral absorption coefficient of signal in ionosphere.

TABLE I-15

Altitude z, km	N_e, cm^{-3}	ν_{eff}, sec^{-1}
60	10	$6.2\cdot10^7$
70	50	$1.6\cdot10^7$
80	$6\cdot10^2$	$3.4\cdot10^6$
90	$6.5\cdot10^3$	$6.7\cdot10^5$
100	$1.5\cdot10^4$	$1.5\cdot10^4$
110	$3.2\cdot10^4$	$3.5\cdot10^4$
120	$1\cdot10^5$	$1\cdot10^4$
150	$1.5\cdot10^5$	$2\cdot10^3$
200	$2.5\cdot10^5$	$8\cdot10^2$
250	$4\cdot10^5$	$6\cdot10^2$
300	$6\cdot10^5$	$5\cdot10^2$
400	$1.1\cdot10^6$	$1.2\cdot10^3$
500	$8\cdot10^5$	$6.9\cdot10^2$
600	$4\cdot10^5$	$2.8\cdot10^2$
800	$1.4\cdot10^5$	$5.6\cdot10^1$
1,000	$5\cdot10^4$	$2.1\cdot10^1$

Calculations show that the value of $\int_{z_0}^{z} N_e \nu_{eff} dz$ is equal to $0.4\cdot10^{17}$ cm^{-2} sec^{-1}.

The ionospheric absorption margin coefficient $\Gamma^{(1)}$ is equal to

$$\Gamma^{(1)} = e^{\mu}. \tag{I-101}$$

With operation at frequency f = 100 Mcps the coefficient $\Gamma^1 = 1.01$. Consequently, with operation at frequencies above 100 Mcps the ionospheric absorption can be neglected.

Fluctuations of UHF Signal Amplitude on Passage Through the Ionosphere. The study of the reflection of radio waves from the ionosphere has shown that the ionosphere almost always consists of small ionized nonhomogeneities /111 which are in a state of rapid alteration, since the lifetime of each of them is short, particularly in the higher regions of the atmosphere where the diffusion increases markedly (ref. 5).

The presence of nonhomogeneities in the ionosphere leads to the electron concentration $N_e(z)$ at level z, being a random function of time t and varying about its mean value $N_{e_0}(z)$

$$N_e(z, t) = N_{e_0}(z) - \Delta N_e(t).$$

Since the wave refraction index n(z) depends on $N_e(z, t)$, it also oscillates about its mean value n_0

$$n(t) = n_0 + \Delta n(t).$$

Limiting ourselves to the consideration of the case of passage of UHF-band radio waves through the ionosphere, for which

$$\left.\begin{aligned} n^2(z) &\approx 1 - \frac{4\pi N_e(z) e^2}{m\omega^2}, \\ n(z) &\approx 1, \end{aligned}\right\} \qquad \text{(I-102)}$$

we obtain the following equation for $\Delta n(t)$ /112

$$\Delta n(t) \approx \frac{4\pi e^2 \Delta N_e(t)}{2m\omega^2}, \qquad \text{(I-103)}$$

where e, m are the electron charge and mass;
ω is the cyclic frequency of the radio wave.

With the propagation of radio waves in this medium we observe fluctuations of the amplitude as the result of scattering of the waves on the nonhomogeneities and the superpositioning of the scattered waves on the primary wave. Let us establish the connection between the fluctuations of the amplitude of the wave A(t) and the fluctuations of the refraction index n(t), for which we shall make use of the results obtained in reference 55, in which it is presumed that the random nonhomogeneities are only in the right semispace (z > 0), and that there are no random nonhomogeneities in the left semispace (z < 0) (fig. I-25). The plane electromagnetic wave

$$\bar{E} = E_0 \exp[-i(\omega t - \kappa z)]$$

propagates from a homogeneous medium into a nonhomogeneous medium.

With normal incidence of the wave on the ionospheric layer (right semi-space), and neglecting the effect of the Earth's magnetic field (which can be done, since $f \gg f_H$, where f_H is the electron gyromagnetic frequency), the equation for the electric field has the form

$$\frac{d^2E}{d^2z} + \kappa^2\varepsilon'(\omega, z)E = 0, \tag{I-104}$$

where E is either of the components E_x or E_y;

$k = \omega n(z)/c$ is the wave number in the medium;
c is the velocity of light;
$\varepsilon'(\omega, z)$ is the dielectric permeability of the medium, in the case of high frequencies equal to $n^2(z)$.

It is shown in reference 55 that equation (I-104) is solved relatively easily with satisfaction of two conditions. /113

(1) smallness of wavelength scattering;
(2) smallness of angle of deviation of scattered ray from initial direction.

We shall show that in the present case both conditions are satisfied. The practicability of the first condition is obvious: the condition of smallness of the wavelength scattering is always satisfied in the case of a weakly nonhomogeneous medium ($\Delta n(t) \ll 1$), which is true for the ionosphere for the frequency range in question.

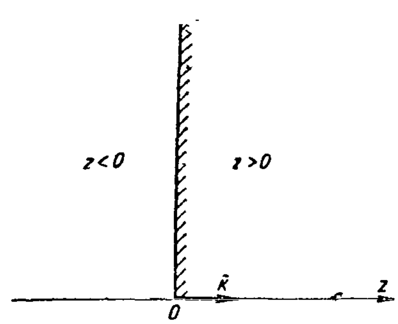


Figure I-25. Coordinate system for calculation of fluctuations of signal amplitude during passage through nonuniform medium.

Let us prove the practicability of the second condition. It is known that the primary scattering of radio waves takes place in the F_2 region of the ionosphere. Here the most probable dimensions of the nonhomogeneities are a = 200 m (ref. 5). In the case of the radio waves of the UHF band $a/\lambda \gg 1$, and these nonuniformities can be considered as large-scale. Since large-scale nonuniformities create sharply directional radiation, the condition of smallness of the angle of deviation of the scattered ray from the original direction is satisfied for them.

With satisfaction of both conditions and on the assumption that the medium is isotropic and that its correlation coefficient is written in the form

$$\rho = \exp\left[-\frac{x^2+y^2+z^2}{a^2}\right] \tag{I-105}$$

(a is the correlation radius, of the same order of magnitude as the scale of the nonuniformities), for the mean square of the logarithm of the ratio of the instantaneous amplitude A(t) to the amplitude of the mirror component A_0 we obtain equation

$$\overline{\left[\ln\frac{A(t)}{A_0}\right]^2} = \frac{\sqrt{\pi}}{2}\overline{\Delta n^2}\kappa^2 aL\left[1-\frac{1}{D}\operatorname{arc\,tg} D\right], \tag{I-106}$$

where

$$D = \frac{4L\lambda}{2\pi a^2};$$

L is the equivalent length of the propagation path in the scattering volume.

Substituting in place of $\overline{\Delta n^2}$ its expression in terms of $\overline{\Delta N_e^2}$, /114

$$\overline{\Delta n^2} = \left(\frac{4\pi e^2}{2m\omega^2}\right)^2 \overline{\Delta N_e^2}$$

and replacing $\overline{\Delta N_e^2}$ by its expression from the relation

$$\sigma = \sqrt{\overline{\left(\frac{\Delta N_e}{N_{e_0}}\right)^2}},$$

we rewrite (I-106) in the form

$$\overline{\left[\ln \frac{A(t)}{A_0}\right]^2} = \frac{\sqrt{\pi}}{2} aL \frac{e^4 \lambda^2 \sigma^2 N_{e_0}^2}{m^2 c^4} \left[1 - \frac{1}{D} \operatorname{arc\,tg} D\right]. \qquad \text{(I-107)}$$

Since in the ionosphere there is a pronounced variation of the electron concentration with height z, both values L and N_{e_0} are indeterminate.

Let us proceed as follows: let us assume that the scattering volume is a "slab" of ionization with the lower boundary z_1, the upper boundary z_2 and the constant value of the electron concentration

$$N_{e_0}(z_1,\ z_2) = \frac{\int_{z_1}^{z_2} N_e(z)\,dz}{z_1 - z_2}. \qquad \text{(I-108)}$$

Let us carry out the calculation of $\overline{[\ln A(t)/A_0]^2}$ for various combinations of z_1 and z_2 (with a 50 km step), and let us select as z_1 and z_2 (consequently, N_{e_0}) those values $z_1 = z^*_1$, $z_2 = z^*_2$ for which the quantity $\overline{[\ln A(t)/A_0]^2}$ reaches a maximum, i.e., we make an estimate of this quantity from this.

Making the calculations for the case of the electron concentration distribution shown in table I-15, we find that these values will be $z^*_1 = 250$ km, $z^*_2 = 600$ km, i.e., L = 350 km. Then $N_{e_0} = 8.5 \cdot 10^5$ cm^{-3}. Taking /115 $\sigma = 10^{-2}$, a = 200 m (ref. 56) and substituting into (I-107) the values of all known quantities, we obtain

$$\overline{\left[\ln \frac{A(t)}{A_0}\right]^2} = 4{,}5 \cdot 10^{-6} \lambda^2 \left[1 - \frac{1}{D} \operatorname{arc.tg} D\right]. \qquad \text{(I-107')}$$

We calculate $\overline{[\ln A(t)/A_0]^2}$ for various λ and summarize the results of the computations in table I-16.

There is basis to presume that the deviations ΔA from the mean value A_0 have a normal distribution in the case of operation at the high frequencies (when the power of the scattered wave is less than the power of the direct wave, ref. 57).

TABLE I-16

λ, cm	10	30	50	75	100	150	200
$\overline{[\ln A(t)/A_0]^2}$	$3.8 \cdot 10^{-5}$	$1.5 \cdot 10^{-3}$	$6.3 \cdot 10^{-3}$	$1.7 \cdot 10^{-2}$	$3.4 \cdot 10^{-2}$	$8.4 \cdot 10^{-2}$	$15.7 \cdot 10^{-2}$

We computed the values of

$$\overline{\left[\ln \frac{A(t)}{A_0}\right]^2} = \overline{y^2}.$$

We need to find the relationship between $\overline{y^2}$ and $\overline{\Delta A^2}$.

It is known that the distribution function of the random quantity $y = f(x)$ is expressed in terms of the distribution function $W(x)$ of the random quantity x as follows

$$W(y) = W(x)[\varphi(y)] \frac{d\varphi}{dy}, \tag{I-109}$$

where $\varphi(y)$ is the inverse function of $f(x)$, expressing x in terms of y. In this case $x = \Delta A$ and

$$f(x) = \ln \frac{x + A_0}{A_0};$$

$$\varphi(y) = A_0(e^y - 1).$$

Consequently /116

$$W(y) = \frac{A_0 e^y}{\sqrt{2\pi}\sqrt{\overline{\Delta A^2}}} e^{-\frac{A_0^2(e^y - 1)}{2\sqrt{\overline{A_0^2}}}} \tag{I-110}$$

By definition

$$\overline{y^2} = \int_{-\infty}^{+\infty} W(y) y^2 \, dy,$$

and then

$$\overline{y^2} = \int_{-\infty}^{+\infty} y^2 \frac{A_0 e^y}{\sqrt{2\pi}\sqrt{\overline{\Delta A^2}}} e^{-\frac{A_0^2(e^y-1)}{2\sqrt{\overline{\Delta A^2}}}} dy. \tag{I-111}$$

Denoting $\frac{A_0}{\sqrt{\overline{\Delta A^2}}}$ by θ and considering θ a parameter, we determine the value of $\overline{y^2}$ for various θ. The results of the computations are summarized in table I-17.

We use the data of the table to construct the graph (fig. I-26) of the variation of $\sqrt{\overline{\left(\frac{\Delta A}{A_0}\right)^2}}$ with $\overline{y^2}$ (interpolation graph). Using the graph and the data of table I-16 we calculate $\sqrt{\overline{\left(\frac{\Delta A}{A_0}\right)^2}} = \sqrt{\overline{x^2}}\frac{1}{A_0}$ for various λ. The results are summarized in table I-18.

The calculated values of $\sqrt{\overline{x^2}}\frac{1}{A_0}$ are used to construct the graph of figure I-27.

/117

It is of interest to make a comparison of the results obtained with experimental values. The data on the magnitude of the fluctuations of the amplitude of the electromagnetic wave, as it passes through the ionosphere, are basically obtained with the aid of observations on the intensity of the radiation of the galactic radio sources ("radio stars"). In particular, reference 58 contains

TABLE I-17

θ	2	3	4	5	6	7	8	9	10	11	12
$\sqrt{\overline{\Delta A^2}}$	$\frac{1}{2}A_0$	$\frac{1}{3}A_0$	$\frac{1}{4}A_0$	$\frac{1}{5}A_0$	$\frac{1}{6}A_0$	$\frac{1}{7}A_0$	$\frac{1}{8}A_0$	$\frac{1}{9}A_0$	$\frac{1}{10}A_0$	$\frac{1}{11}A_0$	$\frac{1}{12}A_0$
$\overline{y^2}\cdot 10^3$	375	175	75,2	46,5	31,4	21,6	16,4	12,4	10,0	8,0	6,0

TABLE I-18

λ, см	10	30	50	75	100	150
$\frac{1}{A_0}\sqrt{\overline{x^2}}$, %	0,5	5	8,5	12,8	17,25	25

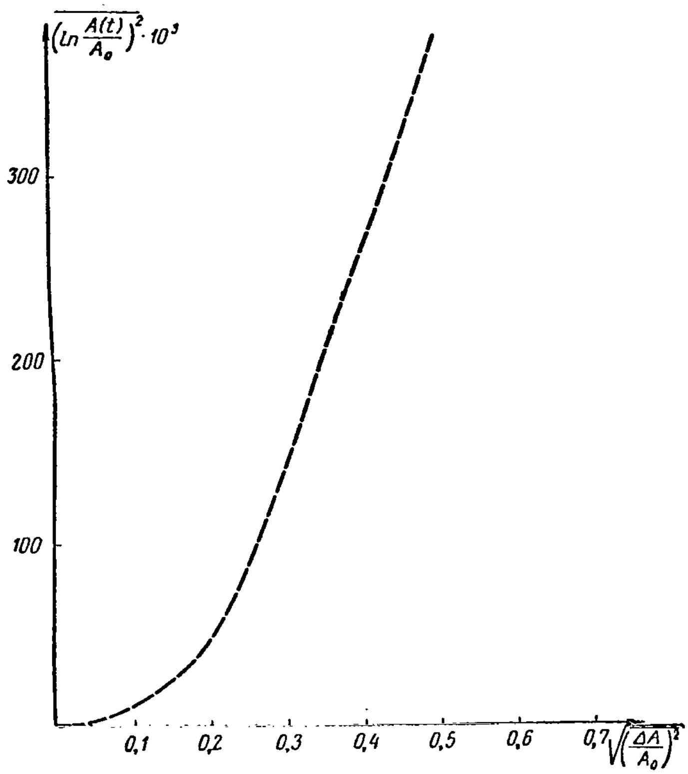


Figure I-26. Mean square of logarithm of fluctuation versus mean fluctuation of amplitude.

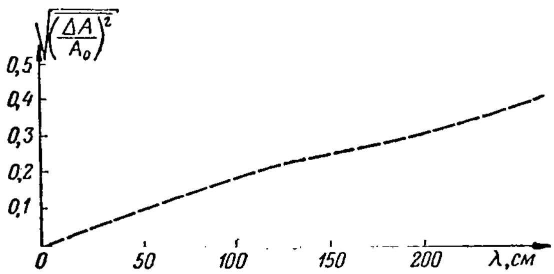


Figure I-27. Variation of average amplitude fluctuation with wavelength.

data on the fluctuation of the intensity of the radiation of the radio source in the constellation Cassiopeia at the frequency f = 79 Mcps (λ = 3.8 m). /118
These data indicate that there are three types of recordings of galactic noise.

(1) records with no fluctuations (small number of records);

(2) records with normal type of fluctuations characterized by the fact that the average amplitude of the fluctuation is equal to 23 percent of the average amplitude of the field (majority of the records);

(3) records during periods of strong disturbances in the terrestrial magnetic field, characterized by high amplitude of the fluctuations. The average amplitude of the fluctuations reaches 67 percent of the intensity of the source (small number of records).

By interpolation we find from figure I-27 that the calculated value of the mean square amplitude of the fluctuations at the wavelength $\lambda = 3.8$ m is equal to about 65-70 percent of the signal intensity. Thus, with strong disturbances of the terrestrial magnetic field (when high "diffusivity" of the F_2 layer is observed), the calculated and experimental values are approximately the same. In reference 59 it is shown that with observation of the intensity of the galactic noise at a frequency of 223 Mcps ($\lambda = 134$ cm) cases were noted when the amplitude of the fluctuations reached 35-50 percent. Although the authors do not indicate the probability of the occurrence of such values of the amplitude /119 of the fluctuations (there is no doubt that these values are higher than the average value), we can still show that in order of magnitude the calculated and experimental values are in quite good agreement. These examples indicate that our results on the preliminary estimate of the intensity of the fading of the radio signal on passage through the nonuniform atmosphere to a certain degree correctly reflect the objective reality and therefore can be used for the selection of the operational wavelength.

Let us determine for the various frequencies the values of the fading (power) margin coefficient $\Gamma^{(2)}$. We calculate $\Gamma^{(2)}$ from the approximate equation

$$\Gamma^{(2)} \approx \left[\frac{1}{1-\sqrt{\overline{\left(\frac{\Delta A}{A_0}\right)^2}}}\right]^2 \qquad \text{(I-112)}$$

The results of the calculation are summarized in table I-19.

These computed values lie on the curve

$$\Gamma^{(2)} = 1 + 0{,}47\,(\lambda - 0{,}07), \qquad \text{(I-112')}$$

with an adequate degree of accuracy, where λ is measured in meters.

Rotation of the wave polarization plane (Faraday effect). With the existence of Earth's magnetic field, the ionosphere is a medium with dual refraction: a radio wave propagating through it is split into two components (normal waves)--the ordinary and the extraordinary. These components propagate

TABLE I-19

λ, cm	10	30	50	75	100	150	200	250	300
Γ(2)	1.015	1.110	1.207	1.315	1.472	1.785	2.040	2.660	3.820

in the ionosphere with differing phase velocities, and therefore after passage through some distance there appears a phase shift between them, which /120 leads to a rotation of the plane of polarization of the overall wave. In the case of a moving radiator (for example, with the installation of the transmitter on board a satellite) there is continuous rotation of the plane of polarization. The signal received by a linearly polarized antenna undergoes polarization fading (ref. 60).

With the following simplifying assumptions

(a) propagation is quasilongitudinal (for high frequencies this is satisfied up to $\alpha = 80^\circ$, where α is the angle between the wave vector and the direction of the terrestrial magnetic field);

(b) $f \gg f_0$; $f \gg f_H$, where f_0 is the plasma frequency in the ionosphere; f_H is the electron gyromagnetic frequency in the ionosphere;

(c) there is no refraction in the ionosphere and the propagation trajectories of both normal waves are the same;

(d) the structure of the ionosphere is layer-wise uniform,

-- the angle of rotation of the wave polarization plane along the propagation path in the ionosphere can be represented in the form

$$\Omega = \frac{K}{f^2}\overline{M}\int_{h_0}^{h(t)} N_e(h)\,dh, \text{ deg.} \qquad \text{(I-113)}$$

Here $K = 2.36 \cdot 10^4$ (Gaussian system of units);
f is the wave frequency, cps;
N_e is the electron concentration in el/cm^2;

$$M = H_0 \cos\theta \sec\delta;$$

H_0 is the intensity of Earth's magnetic field, oe;

h_0 is the height of the lower boundary of the ionosphere, cm;

$h(t)$ is the height of the radiation source, cm;

δ is the zenith angle
θ is the magnetic declination.

At heights near 1,000 km the quantity $M = H_0 \cos\theta \sec\delta$ varies little along the wave propagation path; therefore its averaged value is taken out from under the integral sign. Assuming for simplicity that $\cos\theta \sec\delta = 1$ and that the distribution of the electron concentration $N_e(h)$ in the ionosphere /121 corresponds to the data of table I-15, we obtain

$$\Omega = \frac{2.37 \cdot 10^{17}}{f^2} \text{ deg.} \qquad \text{(I-113')}$$

We calculate Ω for the various frequencies of the UHF band and summarize the results in table I-20.

The expression for the rate of rotation of the polarization plane of the wave is found by means of differentiating (I-113) with respect to time.

An idea of the magnitude of the rate of rotation of the plane of polarization can be given by the data obtained with the aid of calculations using the approximate equation

$$\frac{d\Omega}{dt} = \frac{3.4 \cdot 10^{-3}}{f^2} H_0 N_{e_p} \frac{v}{h} \text{ deg/sec,} \qquad \text{(I-114)}$$

where N_{e_p} is the number of electrons in a column of the ionosphere with a base of 1 m^2;

h is the satellite flight altitude, in m.

As an example let us consider h = 300 km. The satellite flight altitude h = 300 km corresponds to a flight velocity v = 7 km/sec = $7 \cdot 10^3$ m/sec and $N_{e_p} = 3.24 \cdot 10^{17}\ m^{-2}$. In this case

$$\frac{d\Omega}{dt} = \frac{1.8 \cdot 10^6}{f^2} \text{ deg/sec.} \qquad \text{(I-114')}$$

We calculate $d\Omega/dt$ for the various frequencies and summarize the results of the calculation in table I-21. If we assume that in the absence of the /122 ionosphere the polarization vectors of the antennas of the corresponding stations are colinear, the level of a signal with linear polarization received on a linearly polarized antenna varies according to the law

$$A(t) = A_0 \cos [\Omega(t)],$$

i.e., it is subject to deep fading (polarization fading as the result of the Faraday rotation of the polarization plane of the electromagnetic waves). The fading rate $d\Omega/dt$ is inversely proportional to the square of the frequency. To prevent fading it is necessary to use antennas with circular polarization.

Effect of the Troposphere on Radio Wave Propagation

Absorption of Radio Waves in the Troposphere. According to reference 61, the greatest absorption occurs in the oxygen and water vapor of the atmosphere, and also in rain.

Considering that the intensity of the rain falling simultaneously over a large territory will not exceed 40 mm/hr (in this case the height of the rain front will not exceed 2 km), we present some data on the magnitude of the absorption coefficient for radio waves in oxygen and water vapor and also in rain with an intensity of 43 mm/hr (curves a and b, fig. I-28).

It is evident that the magnitude of the absorption coefficient depends on the length of the propagation path in the absorbing medium. Let us determine the equivalent length of the propagation path in an absorping medium for two cases.

(a) absorption in the oxygen and water vapor;
(b) absorption in rain with altitude of rain front 2 km.

The calculation is made for the minimal elevation angle of the receiving antenna $\delta_{min} = 5^\circ$.

For relatively low heights h above the surface of Earth, the barometric equation for the distribution of the concentration of air particles in Earth's atmosphere is valid

$$n = n_0 e^{-\frac{h}{H}}, \tag{I-115}$$

TABLE I-20

f, Mcps	100	200	500	800	1,000	2,000	3,000	5,000
Ω, deg	2,370	595	95	37	24	6	3	1

TABLE I-21

f, Mcps	100	200	500	800	1,000	2,000	3,000	5,000
$d\Omega/dt$, deg/sec	1.7	0.43	$7 \cdot 10^{-2}$	$2.7 \cdot 10^{-2}$	$1.7 \cdot 10^{-2}$	$4.3 \cdot 10^{-3}$	$1.9 \cdot 10^{-3}$	$7 \cdot 10^{-4}$

where n_0 is the particle concentration for h = 0; /123

H is the equivalent height of a uniform atmosphere, equal for air to H = 8.18 km.

The distance d between the points A and B (fig. I-29) is related as follows to the height h of point B above Earth's surface

$$h=\sqrt{d^2+R_0^2+2R_0 d\sin\delta_{min}}-R_0. \tag{I-116}$$

Substituting in place of h its expression in terms of d in the barometric equation and integrating with respect to d in the limits from zero to infinity, we obtain the equation for the determination of the equivalent length of the propagation path (referred to the uniform atmosphere) for the case of absorption in /124 oxygen and water vapor

$$R_{eq}=\int_0^\infty \exp\left[-\frac{d^2+R_0^2+2R_0 d\sin\delta_{min}-R_0}{H}\right]dd. \tag{I-117}$$

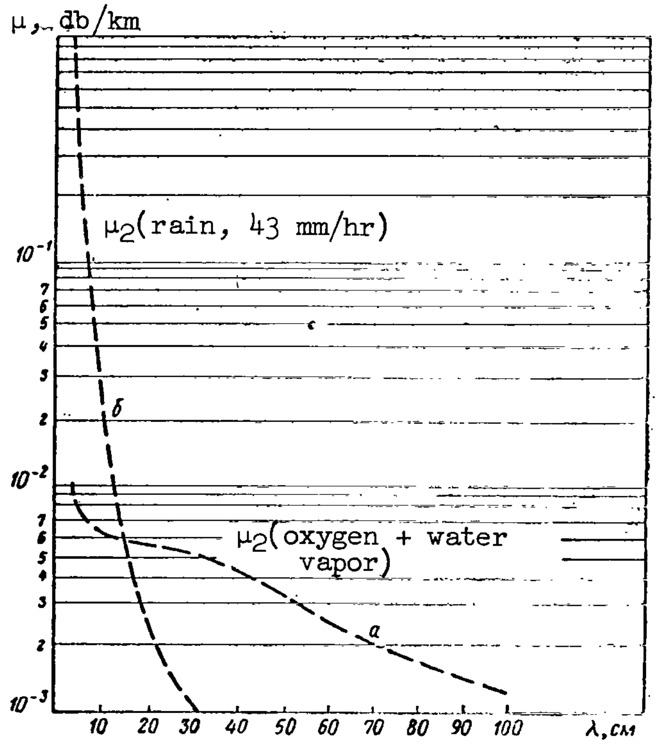


Figure I-28. Variation of signal absorption coefficient in troposphere with wavelength.

Performing the integration with $\delta_{min} = 5^{o}$, we obtain $R_{eq} = 90$ km. In the case of absorption in rain with height of the front h = 2 km, the length of the propagation path with $\delta_{min} = 5^{o}$ is equal to $R'_{eq} = 20$ km. Using curves a and b of figure I-28, we calculate the power margin coefficients $\Gamma_1^{(3)} = \Gamma^{(3)}$; $\Gamma_2^{(3)} = (\Gamma^{(3)})^2$ for compensation of the absorption in oxygen, atmospheric water vapor and rain. The results of the calculations are shown in figure I-30.

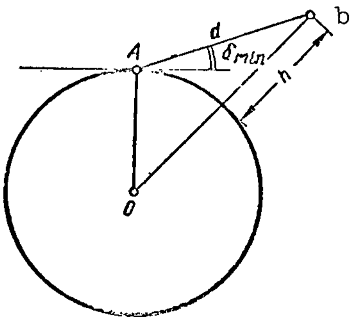


Figure I-29. Geometry of problem on determination of equivalent length of propagation path in uniform atmosphere.

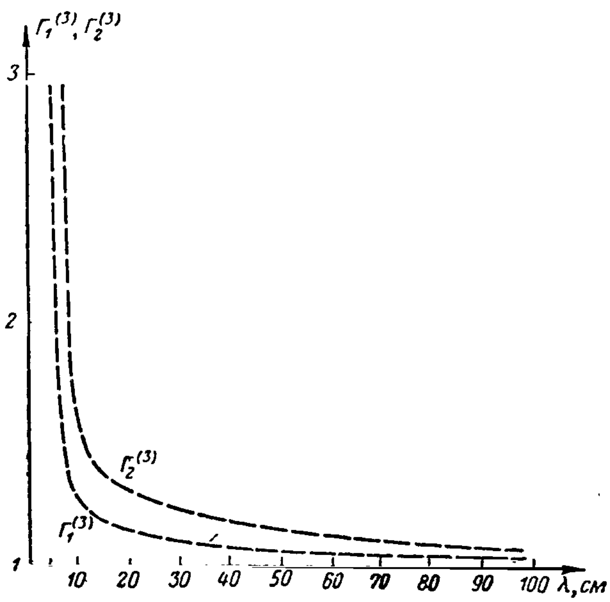


Figure I-30. Margin coefficients $\Gamma_1^{(3)}$, $\Gamma_2^{(3)}$ for tropospheric absorption (active and passive relay, respectively) as function of wavelength.

Refraction of Radio Waves with Passage Through the Troposphere. Refraction of the radio waves (deviation of the path from rectilinearity) is observed both with passage through the troposphere and with passage through the ionosphere. However, the magnitude of the ionospheric refraction is inversely proportional to the square of the frequency (ref. 5), and with operation at the high frequencies its contribution to the overall refraction is not large. The magnitude of the ionospheric refraction is calculated from the equation

$$\Delta\delta_{ion} = -\frac{57.3 \cos \delta}{\sin^3\delta f^2} \text{ [deg]}, \tag{I-118}$$

where f is used in Mcps.

Tropospheric refraction in the radio wavelengths from 3 cm to 3 m does not depend on the wavelength. For elevation angles δ differing somewhat from 0°, the magnitude of the tropospheric refraction is calculated from the equation (ref. 62)

$$\Delta\delta_{trop} = (n-1) \text{ ctg } \delta \text{ [deg]}, \tag{I-119}$$

where n is the tropospheric index of refraction ($n-1 \approx 3\cdot10^{-4}$). The computed values of the angles of tropospheric and ionospheric refraction for different δ at various frequencies are shown in figure I-31.

We can see that with operation at frequencies above 300-400 Mcps the overall refraction is determined primarily by the tropospheric refraction and does not depend on the frequency. With $\delta = 5^\circ$ the refraction is equal to $\Delta\delta = 10'$.

Without accounting for the refraction phenomenon, with programmed tracking of a satellite by a ground-based antenna it is possible to "lose" the satellite, if the width of the ground antenna pattern is less than the refraction angle.

It is known that for good antenna operation it is necessary to have a high degree of accuracy of fabrication of the reflector: the deviations must not exceed $\lambda/16$ (ref. 52). Usually the accuracy of fabrication of the antenna is characterized by the relative fabrication accuracy, equal to $\lambda/16D$ (D is the antenna diameter). At the present time antennas are built with a relative fabrication accuracy which reaches 10^{-4}. With a constant diameter of the ground antenna, this limits the upper band of the operating frequencies. With operation at a frequency critical for the given diameter ($\lambda = 16D\cdot10^{-4}$), the width of the antenna pattern at the half-power points is equal to approximately 10' (antenna surface area utilization coefficient taken equal to 0.5). /126

With operation at the wavelength $\lambda = 30$ cm, the critical diameter of the antenna is equal approximately to $D^* \approx 187$ m; with the use of an antenna with D = 20 m, the critical wavelength is equal approximately to $\lambda^* \approx 3.2$ cm.

Thus, accounting for refraction in the compiling of the program for the antenna drive is necessary with small elevation angles δ and for operation at

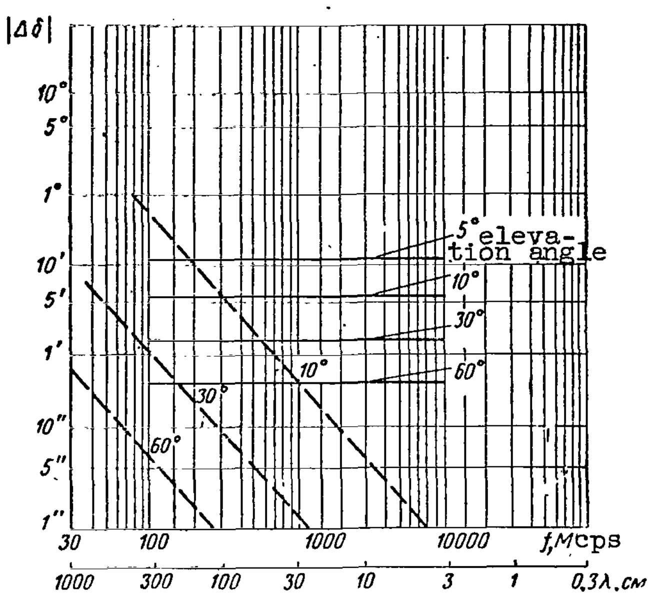


Figure I-31. Variation of tropospheric refraction (solid lines) and ionospheric refraction (dashed) with frequency for various elevation angles.

frequencies close to the critical for the given diameter of the ground antenna D_2. If the operating frequency is above 300-400 Mcps and use is made of an antenna with pattern width exceeding 1°, refractions can be neglected.

In addition to regular refraction due to the nonunity value of the medium index of refraction, which can be compensated by the introduction of a correction, there is also irregular refraction, due to the existence of fluctuation of the medium index of refraction, which cannot be accounted for by the introduction of a constant correction. However, as a result of the relative smallness of the fluctuations of the medium index of refraction (troposphere, ionosphere) the mean square value of the angle of irregular refraction is approximately an order of magnitude smaller than the mean square value of the angle of regular refraction, and the irregular refraction can be neglected with the existing accuracy of antenna fabrication. /127

Margin Coefficient N_1. The margin coefficient N_1 characterizes the losses due to the nonuniformity of the field patterns of the transmitting-receiving antennas, the losses due to the inaccuracy of the pointing of the ground antenna at the AES, the losses in the antenna feeder circuit of the receiving station and the polarization losses. On the assumption that on board the satellite we use an antenna with a conventional radiation pattern (not specialized), the margin for the losses due to the nonuniformity of the pattern should be taken equal to 2.

If for passive relaying use is made of an AES in the form of a metallized sphere, the secondary radiation of the satellite surface will be the same in all directions (isotropic radiator), and there will be no need for a margin for the nonuniformity of the AES antenna pattern. We shall take a margin for 2 for the losses in the antenna feeder circuit of the ground receiving antenna. We also take the margin for inaccuracy of the pointing of the ground antenna at the AES as 2 for the case of both passive and active relaying.

The value of the margin coefficient for the polarization losses depends on the mode of polarization of the ground and onboard antennas. Let us dwell in more detail on the question of the selection of the type of polarization of the onboard and ground antennas.

Let us consider the problem of the determination of the energy transmission coefficient between two antennas with elliptical polarization--ground and airborne.

In the coordinate system, whose x, y coincide with the principal axes of the polarization ellipse of the ground receiving antenna, the elliptically polarized signal radiated by the transmitting antenna (AES antenna), can be represented in the form shown in figure I-32, where angle β is the angle between the semimajor axes of the polarization ellipses of the receiving antenna and the signal emitted by the transmitter. /128

It is known that the elliptically polarized signal can be separated into two circularly polarized signals with opposite directions of rotation, in the general case having differing amplitudes A'_1, A'_2 and different phases φ'_1, φ'_2 (fig. I-33). Let us determine the relationship between A'_1, A'_2, φ'_1, φ'_2 and A_0, B_0, β.

Equating the components E_x, E_y of the incident signal and the circularly polarized waves into which it can be decomposed, we obtain /129

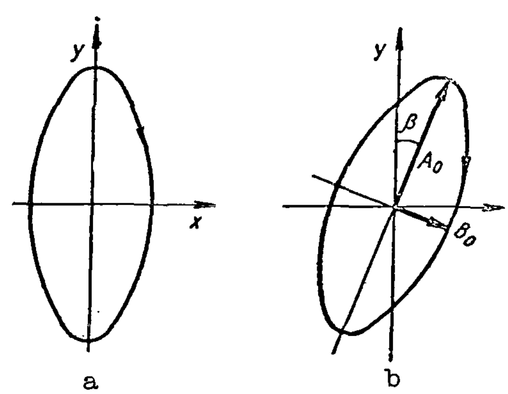


Figure I-32. Polarization ellipses of receiving antenna (a) and signal emitted by transmitter (b).

$$\left.\begin{aligned}&A_0\cos(\omega t+\varphi_0)\cos\beta-B_0\sin(\omega t+\varphi_0)\sin\beta=\\&=A'_1\cos(\omega t+\varphi'_1)+A'_2\cos(\omega t+\varphi'_2),\\&A_0\cos(\omega t+\varphi_0)\sin\beta+B_0\sin(\omega t+\varphi_0)\cos\beta=\\&=A'_1\sin(\omega t+\varphi'_1)-A'_2\sin(\omega t+\varphi'_2).\end{aligned}\right\}\qquad (I-120)$$

Solving the system of equations (I-120) we find

$$\left.\begin{aligned}&\varphi'_1=\varphi_0+\beta,\ \ A'_1=\frac{A_0+B_0}{2}=\frac{A_0}{2}(1+e_1),\\&\varphi'_2=\varphi_0-\beta,\ \ A'_2=\frac{A_0+B_0}{2}=\frac{A_0}{2}(1-e_1),\end{aligned}\right\}\qquad (I-121)$$

where e_1 is the eccentricity of the polarization ellipse of the incident wave (ratio of semimajor and semiminor axes). Thus, the circularly polarized components into which the incident elliptically polarized wave is decomposed can be represented in the form

$$\left.\begin{aligned}&\overline{A'_1}=\bar{i}\frac{A_0}{2}(1+e_1)\cos(\omega t+\varphi_0+\beta)+\\&+\bar{j}\frac{A_0}{2}(1+e_1)\sin(\omega t+\varphi_0+\beta),\\&\overline{A'_2}=\bar{i}\frac{A_0}{2}(1-e_1)\cos(\omega t+\varphi_0-\beta)-\\&-\bar{j}\frac{A_0}{2}(1-e_1)\sin(\omega t+\varphi_0-\beta),\end{aligned}\right|\qquad (I-122)$$

where $\bar{j}$, $\bar{i}$ are unit vectors directed along the axes x, y.

We can show that the receiving antenna with elliptic polarization is equivalent to a system of two circularly polarized antennas A_I, A_{II}, having

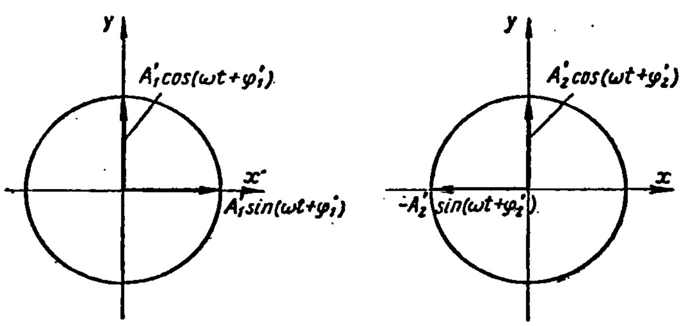


Figure I-33. Circularly polarized signals with right- or left-handed rotation of polarization vector.

opposite directions of rotation and transmission coefficients k_1, k_2 (directional) whose ratio is

$$\frac{k_1}{k_2} = \frac{1 + e_2}{1 - e_2},$$

where e_2 is the eccentricity of the polarization ellipse of the receiving antenna (fig. I-34). /130

The phase shifts of the antennas A_I, A_{II} must be taken to be the same, since the principal axes of the polarization ellipse of the receiving antenna coincide with the coordinate axes y, x.

In turn, the antenna with circular polarization can be represented in the form of two spatially orthogonal linearly polarized antennas, between which the phase shift is plus 90° or minus 90°, depending on the direction of rotation of the polarization circle of the circularly polarized antenna (fig. I-35).

We note that the polarization vectors of the linearly polarized antennas must coincide with the axes x, y of the coordinate system which we have selected.

Thus, the elliptically polarized receiving antenna is equivalent to the antenna system shown in figure I-36. The transmission coefficient between the two antennas with elliptical polarization can be written in the form

$$K = c\,\frac{P_{re}}{P_{tr}}, \tag{I-123}$$

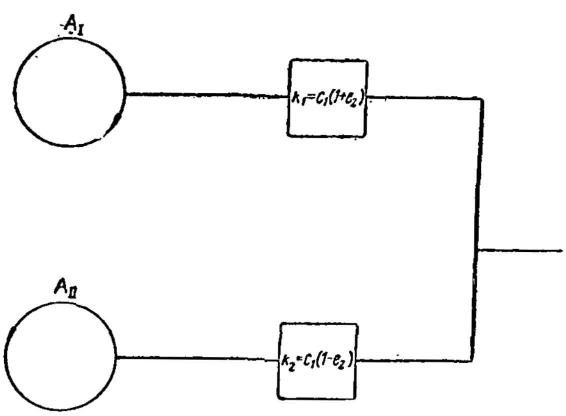


Figure I-34. Equivalent circuit of receiving antenna with elliptical polarization.

where P_{re} is the received power;

P_{tr} is the transmitted power;

c is a constant determined from the normalization condition.

Evidently, on the dipoles A_1, A_2 there are induced the emf's

(a) with arrival of a right-hand polarized wave

$$A'_{\text{I}} = \frac{A_0}{2}(1+e_1)\cos(\omega t+\varphi_0+\beta);$$

$$A'_{\text{II}} = \frac{A_0}{2}(1+e_1)\sin(\omega t+\varphi_0+\beta);$$

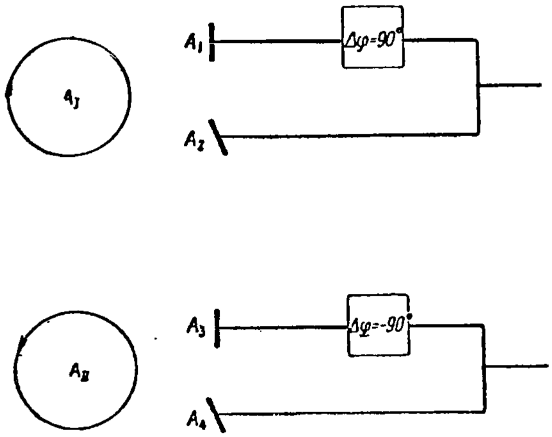


Figure I-35. Equivalent circuit of antenna with circular polarization.

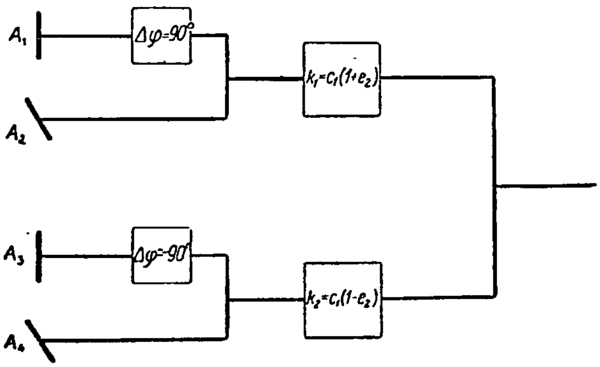


Figure I-36. Complete equivalent circuit of receiving antenna with elliptic polarization.

(b) with arrival of a left-hand polarized wave /132

$$A'_{I}=\frac{A_0}{2}(1-e_1)\cos(\omega t+\varphi_0-\beta),$$

$$A'_{II}=\frac{A_0}{2}(1-e_1)\sin(\omega t+\varphi_0-\beta).$$

After introduction of a 90° phase shift and summing (powers), the resulting signals on the antenna A_1 will be

$$\left.\begin{array}{l} \text{a) } A_I=\frac{A_0}{\sqrt{2}}(1+e_1)\cos(\omega t+\varphi_0+\beta); \\ \text{b) } A_I=0. \end{array}\right\} \quad (I-124)$$

The same emf's are induced on the dipoles A_3, A_4 as on the dipoles A_1, A_2. However, as the result of the -90° phase shift the resulting signals on the antenna A_{II} will be

$$\left.\begin{array}{l} \text{a) } A_{II}=0; \\ \text{b) } A_{II}=\frac{A_0}{\sqrt{2}}(1-e_1)\cos(\omega t+\varphi_0-\beta). \end{array}\right\} \quad (I-125)$$

Assuming the coefficients k_1, k_2 equal to

$$k_1=c_1(1+e_2),\quad k_2=c_1(1-e_2)$$

(where c_1 is some constant number), after addition of the signals (I-124) and (I-125) (powers are summed) we obtain at the output of the receiving antenna the signal

$$\begin{aligned} A_{\Sigma}=\frac{A_0}{2}c_1[&(1+e_1)(1+e_2)\cos(\omega t+\varphi_0+\beta) \\ &+(1-e_1)(1-e_2)\cos(\omega t+\varphi_0-\beta)]. \end{aligned} \quad (I-126)$$

Thus, the received power will be

$$\begin{aligned} P_{re}=c_1^2\frac{A_0^2}{8}[&(1+e_1^2)(1+e_2^2)+(1-e_1^2)^2(1-e_2^2)^2 \\ &+2\cos 2\beta(1-e_1^2)(1-e_2^2)]. \end{aligned} \quad (I-127)$$

It is evident that the transmitted power is /133

$$P_{tr} = \frac{A_0^2}{2} c_2 (1 + e_1^2), \tag{I-128}$$

where c_2 is a constant.

After simple manipulations, we find that the energy transmission coefficient between the two antennas with elliptic polarization K can be represented in the form

$$K = \frac{c}{2}\left[(1 + e_2^2) + \frac{4e_1e_2}{1+e_1^2} + \frac{(1-e_1^2)(1-e_2^2)}{1+e_1^2}\cos 2\beta\right]. \tag{I-129}$$

We find c, considering that K = 1 with $e_1 = e_2$ and $\beta = 0$ (polarization ellipses of transmitting and receiving antennas coincide)

$$c = \frac{1}{1+e_2^2}. \tag{I-130}$$

Consequently, the normed energy transmission coefficient can be finally written in the following form

$$K = \frac{1}{2}\left[1 + \frac{4e_1e_2}{(1+e_1^2)(1+e_2^2)} + \frac{(1-e_1^2)(1-e_2^2)}{(1+e_1^2)(1+e_2^2)}\cos 2\beta\right]. \tag{I-131}$$

Thus, knowing the relative positions and parameters of the polarization ellipses of the onboard and ground antennas, from (I-131) we can find the (power) loss of a given receiving antenna in comparison with the receiving antenna, whose polarization ellipse coincides exactly with the polarization ellipse of the incident signal.

Calculations were made for K with various combinations of e_1, e_2, β, and the results are shown in figure I-37.

Analysis of the curves shows that in the case of communication between moving stations (β changes) the optimal variant is that in which the onboard and ground antennas are circularly polarized, since the energy transmission coefficient does not depend on the angle β (axial symmetry) and has its /136 maximal value ($K = K_{max} = 1$). Consequently, there is no need to introduce a margin for the polarization loss in this case. However, in this case of communication of a ground station with a satellite, there are sometimes obstacles to the implementation of this version of the design of the radio link.

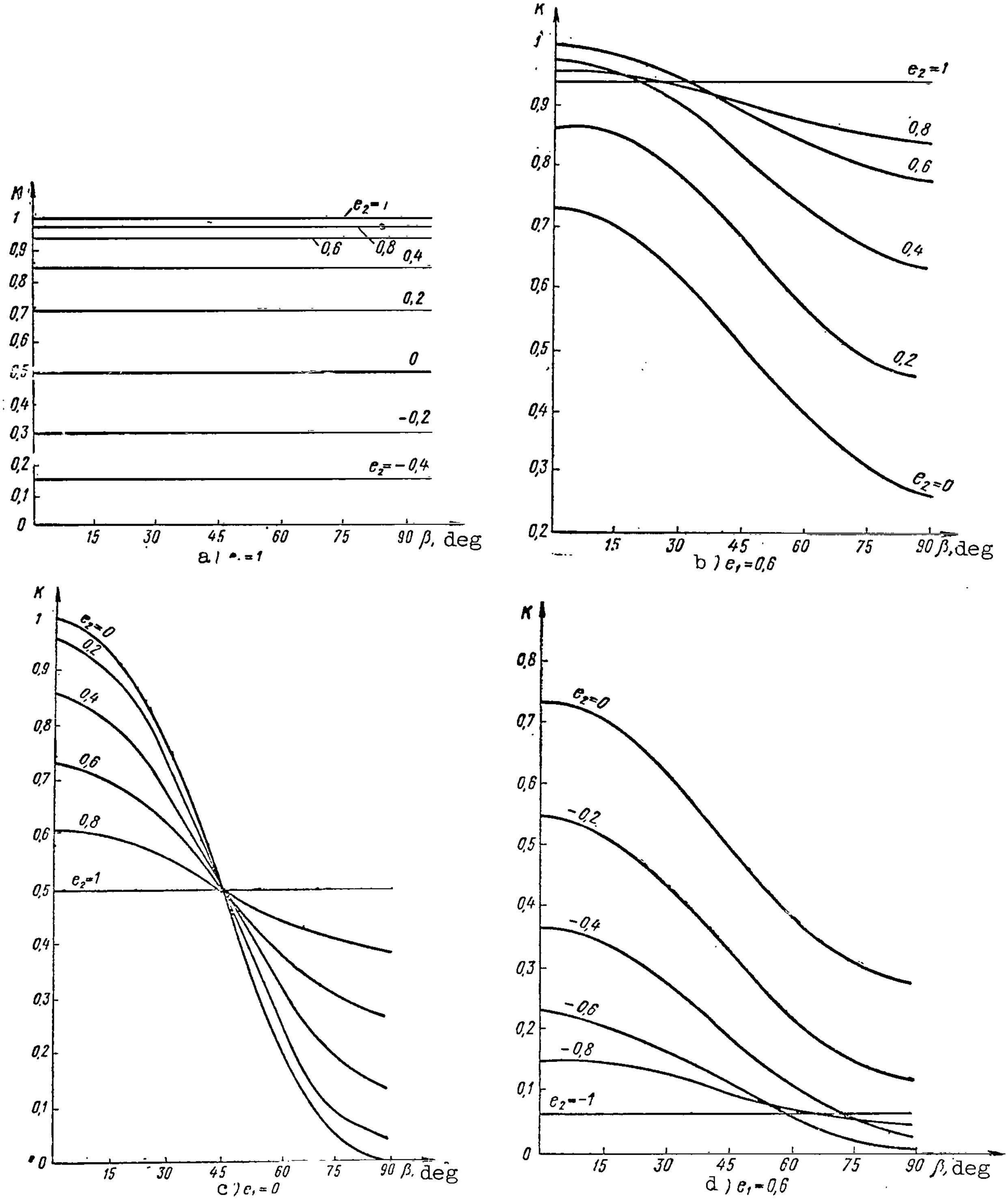


Figure I-37. Energy transmission coefficient between two antennas with elliptic polarization as function of angle β between semimajor axes of polarization ellipses.

There are two factors which prevent the universal implementation of this variant:

(a) the limitation of the width of the field pattern in which circular (or nearly so) polarization of the field can be obtained;

(b) the effect of the variation of the direction of rotation of the circularly polarized wave during its propagation in a magnetically active medium (ionosphere).

We noted that in the case of communications satellites the width of the antenna pattern of the satellite (at the half-power points) must be equal to $\theta = \alpha + 2\gamma$, where α is the angle of coverage of Earth's surface from the satellite, equal to

$$\alpha = 2\arcsin\frac{R_0}{R_0 + h}, \qquad \text{(I-132)}$$

and γ is the correction for instability in orientation of the satellite toward the center of Earth (fig. I-38).

At the present time ground antennas can be fabricated with a maximal width of the pattern in which nearly circular polarization is achieved of up to 100° (ref. 63). If we assume that it is possible to build an onboard antenna for the AES with the same polarization parameters, this corresponds, with $\gamma = \pm 5^\circ$, to a satellite flight altitude h = 2,000 km. With lower satellite flight altitudes θ increases, and circular polarization of the field (one direction of rotation) will not be obtained over the entire width of the radiation pattern, but only in the central portion; closer to the edges of the pattern the polarization becomes increasingly more elliptical, and with very broad patterns, when $\theta \geq 180^\circ$ (this case is realized with a large instability correction γ), it passes through linear and again becomes elliptic, but with opposite direction of rotation (refs. 63 and 52). /137

Analysis of the curves of figure I-37, relating to the case $e_1 = 1$, shows that with different signs of e_1 and e_2 the transmission coefficient K begins to decrease sharply with increase of e_2, and with $e_2 = -1$ it becomes zero. Thus, if an attempt is made to use antennas with circular polarization on the ground and on board the satellite, in the case of relatively low flight altitudes it is possible to have considerable attentuation of the signal on the edges of the antenna pattern, which is extremely undesirable, since it is here that the distance between the corresponding stations is maximal and, consequently, the power consumption will also be maximal.

In the cases of higher satellite flight altitudes, the argument on point (a) naturally loses force, since the provision of an onboard antenna with circular polarization of the field (or close to it) over the entire pattern does not present any difficulty.

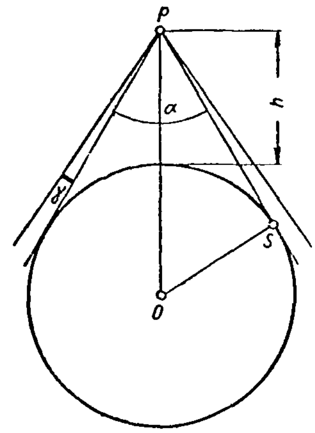


Figure I-38. Geometry of problem on determination of field of view of Earth's surface from satellite.

The only factor preventing the unrestricted simultaneous use of antennas with circular polarization for transmission and reception in this case is the effect of the variation of the direction of rotation of the circularly polarized wave during passage through the ionosphere. Let us consider this phenomenon in more detail.

The eccentricity of the normal waves (ordinary and extraordinary components) at the altitude z in the approximation of geometric optics is written in the form (ref. 64)

$$\frac{E_{y_{1,2}}}{E_{x_{1,2}}} = e_{1,2} = -i\frac{2\sqrt{u}(1-is-v)\cos\alpha}{u\sin^2\alpha \pm \sqrt{u^2\sin^4\alpha + 4u(1-is-v)^2\cos^2\alpha}}, \tag{I-133}$$

where plus in the denominator refers to the ordinary wave (subscript 2), minus to the extraordinary wave (1). Notation is the same as previously used (see "Absorption of Radio Waves in the Ionosphere"). /138

In the case of relaying of signals via AES, the operating frequency must be higher than the critical frequency of the layer f_0. Consequently, the condition $f \gg f_0$, $f \gg f_H$, $f \gg \nu_{eff}$ is satisfied, and with a high degree of accuracy equation (I-133) can be rewritten in the form

$$e_{1,2} \approx -i\frac{2\sqrt{u}\cos\alpha}{u\sin^2\alpha \pm \sqrt{u^2\sin^4\alpha + 4u\cos^2\alpha}}. \tag{I-134}$$

In the case of angles α which are not close to 90°, the condition

$$4u\cos^2\alpha \gg u^2\sin^4\alpha \tag{*}$$

is satisfied, and the equation is rewritten

$$e_{1,2} \approx \pm i \frac{1}{1 \pm \frac{\sqrt{u}\sin^2\alpha}{2\cos\alpha}}, \tag{I-134'}$$

i.e., propagation with angles α which are not close to 90° can be treated as quasilongitudinal propagation; here the polarization of the normal waves is close to circular. For example, at a frequency f = 1,000 Mcps the ellipticity coefficient of the ordinary wave e_2 is equal to 1.1 only with $\alpha = 89^{\circ}34'$. Consequently, the condition of quasilongitudinal propagation is satisfied over practically the entire range of variation of the angles α = 0 to 90°, with the exception of a small region of this range directly adjacent to the value $\alpha = 90^{\circ}$.

Thus, during the entry into the ionosphere of a circularly polarized high-frequency signal there is practically never any appreciable intensity of parasitic component generated with oppositely directed rotation, and the electromagnetic wave propagates in the ionosphere in the form of one of the magneto-ionic components (ordinary or extraordinary), i.e., the transmitting device which radiates radio waves with circular polarization is "matched" with the "entry" into the ionosphere.

We can judge the degree of matching by the following numerical example: at the frequency f = 1,000 Mcps with $\alpha = 89^{\circ}34'$ the ratio of the intensities of the normal waves is equal to n = 441 (the generation of the magnetic components proceeds in this same proportion with incidence onto the ionospheric layer /139 of a circularly polarized wave), and the wave with a direction of rotation equal to the direction of rotation of the incident wave predominates. Consequently, a wave incident on the ionospheric layer propagates in the ionosphere in the form of one of the magneto-ionic components, and its polarization changes in accordance with expression (I-133) as a result of the variation of the medium parameters (N_e, ν_{eff}) and the angle between the direction of propagation of the wave and the magnetic field of Earth α. In this case the variations of the angle α have the strongest influence on the polarization properties of the signals; the variations of the medium parameters make such a small contribution at the high frequencies that their effect can be neglected. A particularly pronounced variation of the polarization properties of the signal with the angle α is observed in the neighborhood of $\alpha = 90^{\circ}$.

Let us consider the behavior of the signal in the vicinity of the point $\alpha = 90^{\circ}$, when the condition

$$4u\cos^2\alpha \ll u^2\sin^4\alpha. \qquad (**)$$

is satisfied. With satisfaction of the condition (**) equation (I-134) is re-written in the form

$$e_{1,2} \approx -i\frac{2\sqrt{u}\cos\alpha}{u\sin^2\alpha \pm u\sin^2\alpha\left(1+\frac{2\cos^2\alpha}{u_c\sin^4\alpha}\right)}. \qquad \text{(I-134")}$$

Replacing α by $\alpha = 90^\circ - \chi$ we obtain

for the ordinary wave

$$e_2 \approx i\frac{\sqrt{u}}{\sin\chi};$$

for the extraordinary wave

$$e_1 \approx -i\frac{\sin\chi}{\sqrt{v}}.$$

Thus, in the vicinity of $\alpha = 90^\circ$ ($\chi = 0^\circ$) the normal waves are elliptically polarized and their polarization ellipses are conjugate ($e_1 e_2 = 1$). The sign of the ellipticity coefficient $e_{1,2}$, which formally defines the direction of rotation of the polarization vector, depends on the sign of χ, i.e., the sign of $e_{1,2}$ changes with transition of χ through 0° (α through 90°).

This analysis makes it possible to draw the conclusion that if geo- /140
metrical optics are applicable, the signal trajectory passes through the point at which the wave normal is perpendicular to the line of force of Earth's magnetic field, and if this point lies in the ionosphere, then the direction of rotation of the polarization vector of the circularly polarized electromagnetic wave propagating through the ionosphere will be reversed.

It is evident that this phenomenon must lead to a significant attenuation of the signal as the result of the "mismatching" of the polarization of the transmitting and the receiving antennas (fig. I-37).

Let us see how frequently this condition is satisfied in practice. According to reference 65 the equation of the line of force of Earth's magnetic field in the polar coordinates r, φ (the coordinate plane includes the line of force, figure I-39) can with adequate accuracy be written as

$$r = \frac{R_0\cos^2\varphi}{\cos^2\varphi_0}, \qquad \text{(I-135)}$$

where R_0 is Earth's radius; /141

φ is the geomagnetic latitude, measured from the magnetic equator;

φ' is the geomagnetic latitude of the point of observation;

SP is the radio wave trajectory;

P' is the point of intersection of the wave trajectory with the line emerging from the point with $\varphi = \varphi_0$.

It can be shown that the condition of perpendicularity of the vectors of the wave normal and Earth's magnetic field is written in the form

$$\frac{\cos^2 \varphi}{\cos^2 \varphi_0} \sin 2\rho = \cos(\varphi - \varphi') \sin 2\varphi + \cos^2 \varphi \cdot \sin(\varphi - \varphi'). \qquad \text{(I-136)}$$

In order to judge the probability of the fulfillment of the conditions required for the occurrence of the effect in question, it is necessary to determine for various values of φ' the range of the interval of the angles Δ in which the condition $\alpha = 90^\circ$ is satisfied on the portion of the wave trajectory lying in the ionosphere.

It is evident that we should take as the coordinate of one end of the interval (φ_1) the coordinate of the point P'_1 which lies on the lower boundary of the ionosphere.

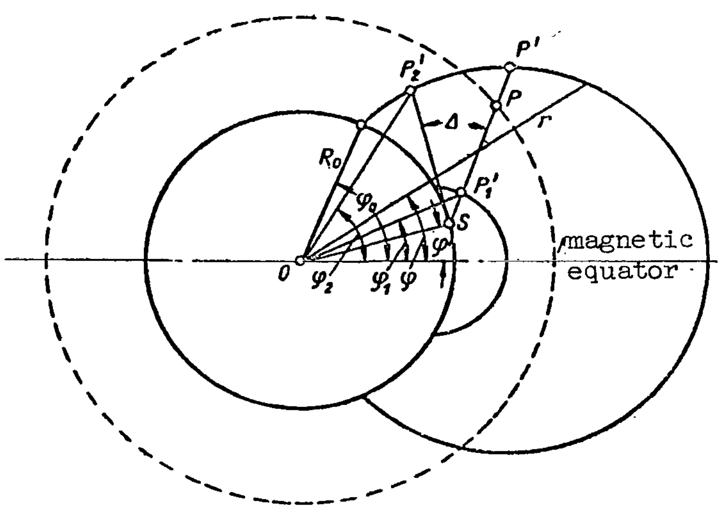


Figure I-39. Geometry of problem on determination of "dangerous" interval of angles.

Taking the height of the lower boundary of the ionosphere h_1 = 100 km and requiring the satisfaction of $\alpha = 90^\circ$ at point P'_1, we find the equation for the determination of φ_1

$$\frac{R_0 + h}{R_0} \sin 2\varphi_1 = \cos(\varphi_1 - \varphi') \sin 2\varphi_1 + \cos^2 \varphi_1 \sin(\varphi_1 - \varphi'). \qquad \text{(I-137)}$$

As the coordinate of the other end of the "dangerous" interval let us take the coordinate φ_2 of point P'_2, which is the point of intersection of the tangent to Earth's surface at the point of observation S and the corresponding magnetic line of force. Considering that the radius vector of the point P'_2 satisfies the relation

$$r_2 = \frac{R_0}{\cos(\varphi_2 - \varphi')}, \qquad \text{(I-138)}$$

we find the equation for the determination of φ_2

$$\cos^2 \varphi_2 = \operatorname{tg}(\varphi_2 - \varphi'). \qquad \text{(I-139)}$$

Solving (I-137), (I-138) and (I-139) we find φ_1 and φ_2 and from them /142 determine h_2 and Δ for the various values of φ'. The results of the calculation are shown in figures I-40 and I-41.

Analysis of the curve of figure I-41 shows that the width Δ of the "dangerous" interval of the angles diminishes with movement of the observation point from the magnetic equator toward the magnetic poles. Consequently, the effect of the variation of the direction of rotation of the circularly polarized wave must appear most frequently with location of the observation point in the vicinity of the magnetic equator.

In the case of communication of a ground station with a satellite, the width of the "dangerous" interval Δ will depend on the shape (circular or elliptic) and the parameters of the orbit (inclination, apogee height, perigee, etc.) and the time t.

It is evident that the width of the "dangerous" interval of angles Δ will be maximal when the projection of the satellite track on Earth's surface is close to one of the magnetic meridians. In this case we should take as Δ the calculated values which we have presented in (fig. I-41).

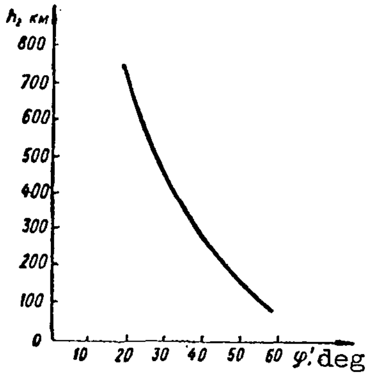


Figure I-40. Variation of altitude of point of signal trajectory with $\alpha = 90^{o}$ with geomagnetic latitude of point of observation φ'.

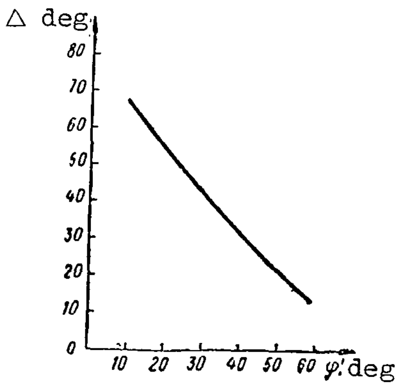


Figure I-41. Variation of width of "dangerous" interval of angles Δ with geomagnetic latitude of observation point φ'.

Similar calculations show that during communication with a satellite in an equatorial (or nearly so) orbit the conditions for the occurrence of the effect of the variation of the direction of rotation of the polarization /143 vector are absent, since the points of the signal trajectory with $\alpha = 90^{o}$ lie beyond the limits of the ionosphere and outside the orbit.

We noted that a necessary condition for the occurrence of the effect of the variation of the direction of rotation of the circularly polarized signal during passage through the ionosphere is the condition of the applicability of geometric optics. If the condition of applicability of geometric optics is not satisfied, there occurs "polarization degeneration," which means that the polarization of the electromagnetic signal is not altered on passage through the ionosphere. In reference 64 the condition of applicability of geometric optics is written in the form

$$|\Psi_1^2| \leqslant \left| \frac{\pi^2 f^2}{c^2}(n_2 - n_1)^2 - \frac{i}{2}\frac{2\pi f}{c}\frac{d}{dz}(n_2 - n_1) \right| = |\Psi_2^2|, \tag{I-140}$$

where

$$\Psi_1 = \frac{i}{4}\frac{d}{dz}\left\{ \ln \left[\frac{1 - v - is + i\dfrac{\sqrt{u}\sin^2\alpha}{2\cos\alpha}}{1 - v - is - i\dfrac{\sqrt{u}\sin^2\alpha}{2\cos\alpha}} \right] \right\};$$

$$n_2 - n_1 = \frac{1}{2} v \sqrt{u^2 \sin^4\alpha + 4u\cos^2\alpha},$$

n_2-n_1 is the difference of the refraction indices of the ordinary and extraordinary waves;

c is the velocity of light.

Note. As before we consider $f \gg f_0$, $f \gg f_H$, $f \gg \nu_{eff}$.

We are interested in the practicability of the applicability of the condition of geometric optics at the point with $\alpha = 90^\circ$. Assuming that $\alpha = 90^\circ$, we obtain

$$\left.\begin{aligned} |\Psi_1^2| &= \left(\frac{d\alpha}{dz}\right)^2 \frac{1}{u}, \\ |\Psi_2^2| &= \left| \frac{\pi^2 f^2 u^2 v^2}{4c^2} - i\frac{\pi f}{c}\left(\frac{du}{dz}\frac{v}{2} + \frac{dv}{dz}\frac{u}{2}\right) \right|. \end{aligned}\right\} \tag{I-140'}$$

We made computations of $d\alpha/dz$ at points of the trajectory with $\alpha = 90^\circ$ /144 for different trajectories and found that $d\alpha/dz \geq 2\cdot10^{-9}$ cm^{-1}.

Since in Earth's ionosphere, from the data of reference 5, $u(z) \leq 2\cdot10^{-12} f^{-2}$, $N_e(z) \leq 2\cdot10^6$ cm^{-3}, $|dN_e/dz| \leq 10^{-1}$ cm^{-4}, it is not difficult to see that

$$\left.\begin{aligned} |\Psi_1^2| \geqslant |\Psi_1^2|_{\mathrm{min}} &= 2\cdot10^{-30} f^2 \ \text{см}^{-2}, \\ |\Psi_2^2| \leqslant |\Psi_2^2|_{\mathrm{max}} &= 10^9 f^{-3}\sqrt{1 + 10^{46} f^{-6}} \ \text{см}^{-2} \end{aligned}\right\} \tag{I-140''}$$

(frequency f in cps).

We calculate the ratio $\frac{|\Psi_2^2|\ max}{|\Psi_1^2|\ min}$ for various values of the frequency and plot the results in figure I-42.

Analysis of the curve shows that the geometric optics approximation is valid at the point with $\alpha = 90^{\circ}$ only for frequencies not exceeding 40 Mcps. The range from 40 to 100 Mcps is intermediate; it separates the range of applicability of geometric optics (normal waves are independent) and the range in which the quasi-isotropic approximation is applicable for the consideration of the polarization properties of the signal (polarization of the signal does not change with passage through the point with $\alpha = 90^{\circ}$). The quasi-isotropic approximation is valid for frequencies above 100 Mcps.

This analysis makes it possible to draw the conclusion that in the case of operation at frequencies above 100 Mcps the conditions are absent for the

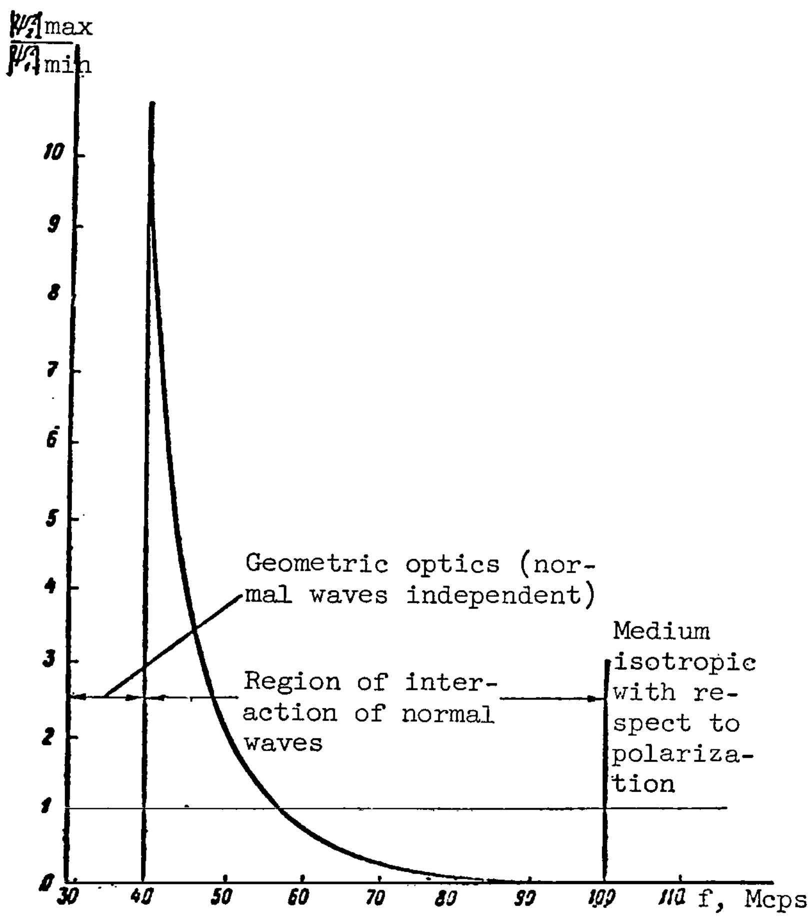


Figure I-42. Determination of limits of acceptability of geometric optics for analysis of signal polarization.

manifestation of the effect leading to a change of the direction of rotation of the circularly polarized signal with passage through Earth's ionosphere.

Consequently, with operation at frequencies above 100 Mcps, in the case of the medium and high orbits ($h > 2,000$ km), the optimal version of the design of the radio link is by use of circularly polarized antennas on board and on the ground (ref. 125). In the case of low orbits ($h < 1,000$ km), when the width of the antenna pattern required for complete coverage of Earth's surface is so large that with the present state of affairs it is not possible to fabricate an antenna with circular polarization of the field (of one sign) over the entire width of the pattern, an antenna with a linearly polarized field should be used on the satellite, and one with a circularly polarized field on the ground. /145 A disadvantage of this approach to the design of the radio link is the additional power loss of a factor of 2 (polarization losses) because of the mismatching of the polarizations of the onboard and ground antennas (ref. 108).

In our examples of the solution of various problems on the provision of ground-to-ground communications using relay via AES, we consider satellites at medium and high altitudes. Therefore, hereafter we shall assume that antennas with circular polarization are used on board and on the ground.

It is obvious that purely circular polarization of the field is obtained only in one direction. In other directions the field polarization is elliptical. However, according to reference 66, in the conventional antennas with elliptical polarization (spiral antennas) the eccentricity does not exceed 2 over the entire width of the field pattern (at the half-power points). Assuming that the eccentricity of the receiving antenna on the ground is ap- /146 proximately equal to 1, we find that the minimal value of the energy transmission coefficient is equal to 0.9. Consequently, in the case of the use of antennas with circular polarization on board and on the ground, the polarization (power) loss margin is only a factor of 1.1.

Summing up all that has been said, we can draw the conclusion that in the case of active relaying the margin coefficient N_1 should be taken equal to 8.8.

With passive relaying we use an antenna with linear polarization for the transmitter and an antenna with circular polarization for the receiver. In this case the polarization loss margin is 2. Consequently, in the case of passive relaying the margin coefficient N_1 should be taken equal to 16. With increase

of the pointing accuracy of the receiving (active relay) or of the receiving and transmitting (passive relay) antennas, power gains of factors of 2 and 4, respectively, for the active and passive relay are possible (margin coefficient N_1 becomes equal to 4.4 and 4, respectively).

Choice of Optimal Wavelength. The analysis presented permits making the selection of the optimal operating wavelength for all radio links mentioned above.

Let us determine $P_t/P_{t_{min}}$ for operation in the various modes for all orbits which we have considered ($h = 2,000$ km, $h = 4,800$ km, $h_a = 20,000$ km,

$h = 36{,}000$ km) and for the following values of the effective self-noise temperatures T_0: 50, 100, 300, 600, 1,500°K. We plot the results of the calculations in figure I-43 (a-e) for active relay and in figure I-44 (a, b) for passive relay. In the case of passive relay, for the indicated values of the noise temperature the curves coincide in the scale of the figure.

We find λ_{opt} and $\Delta\lambda$ for the various combinations of parameters and all orbits (the range of quasi-optimal wavelengths $\Delta\lambda$ is determined from the condition of power increase by a factor of 2) and present the results for the active relay summarized in tables I-22 to I-25.

For the same orbits and the same values of T_0 as used in the cases considered for active relay, we find in all cases of passive relay transmission of FM television or PTM telegraphy, $\lambda = 8$ cm, $\Delta\lambda = 5$ to 12.5 cm, and for /164 the transmission of telephony, $\lambda = 8$ cm, $\Delta\lambda = 5.5$ to 17.4 cm.

Let us carry out the calculation for the required minimal radiation power of the onboard (in the case of active relay) and ground (in the case of passive relay) transmitters $P_{t_{min}}$ for various satellite flight altitudes and various modes of modulation. Here we shall make the calculations in the case of active relay for various values of accuracy of stabilization of the satellite on the assumption that as the ground antenna (receiving in the case of active relay, transmitting and receiving in the case of passive relay) there is used an antenna of the "parabolic mirror" type with paraboloid aperture diameter $D_2 = D_{21} = D_{22} = 20$ m; in the case of passive relay we make the calculations for various values of the effective scattering area of the satellite S_0. If the satellite is in the form of a sphere with metallized coating, it is an isotropic radiator with scattering area $S = \pi D_0^2/4$, where D_0 is the sphere diameter.

The computed values of $P_{t_{min}}$ are plotted in figure I-45 for active relay and figure I-46 for passive relay.

Knowing the minimal value of the required radiation power $P_{t_{min}}$ of the onboard and ground transmitters and using the graph of the relation $P_t/P_{t_{min}}$ (λ) (figs. I-43 and I-44), we can determine the required radiating power P_t for operation in any portion of the range of quasi-optimal wavelengths.

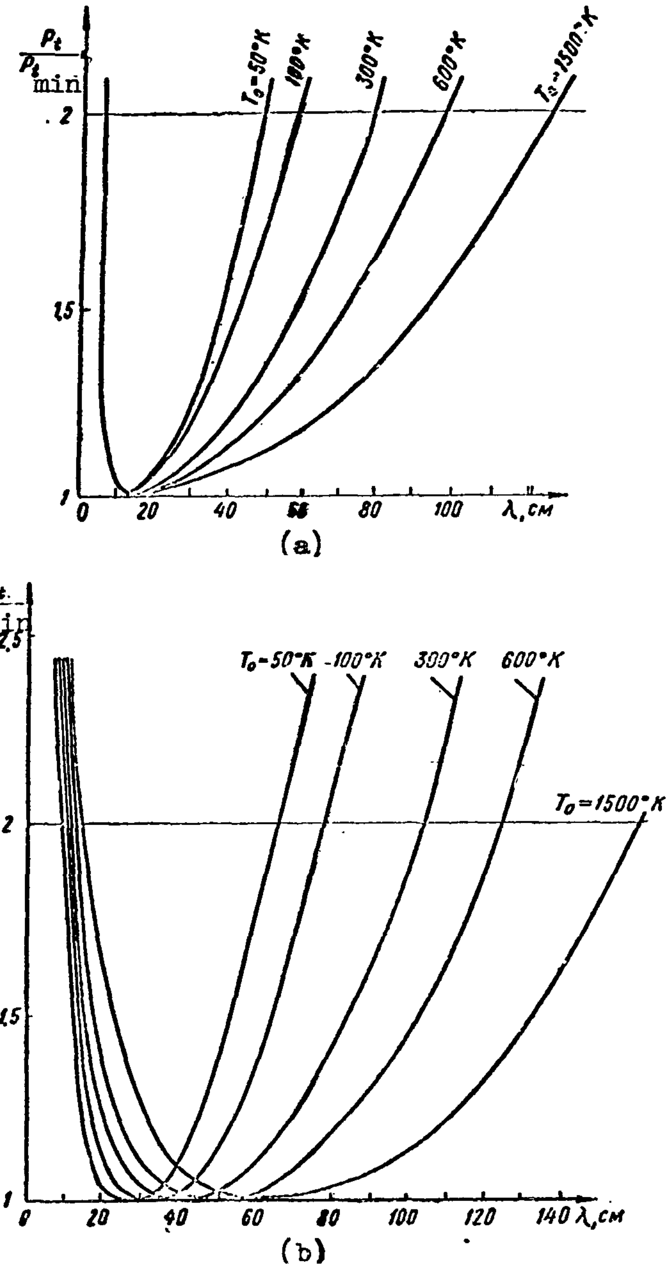
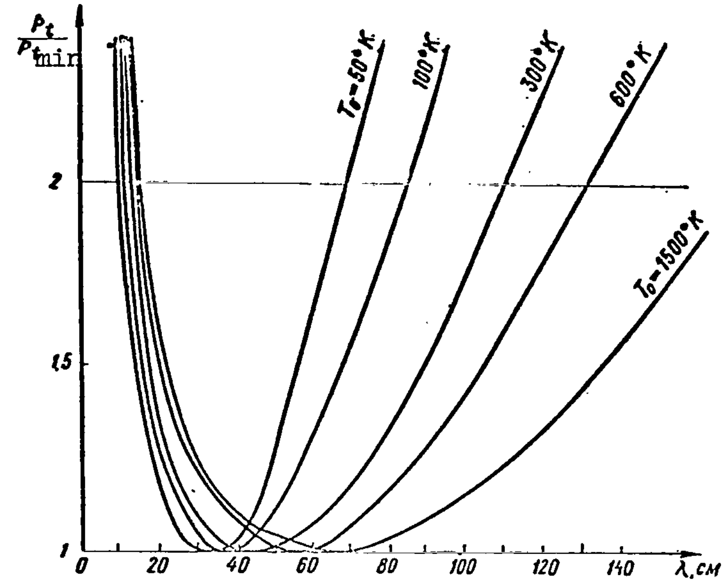
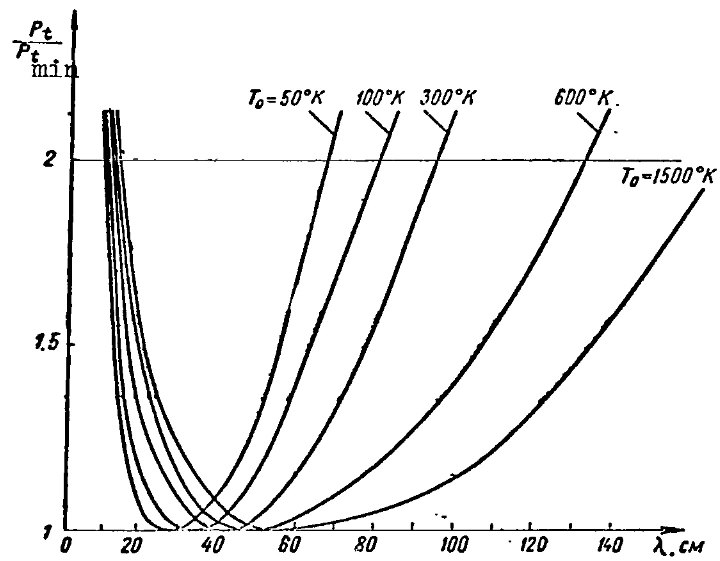


Figure I-43. Dependence of $P_t/P_{t_{min}}$ on wavelength for active relay with $\nu_{des} = 10^{-4}$ for various orbital altitudes. a, FM television (multichannel telephony), PTM telegraphy, all altitudes; b, FM telephony, h = 2,000 km; c, FM telephony, h = 4,800 km; d, FM telephony, h = 20,000 km; e, FM telephony, h = 36,000 km.



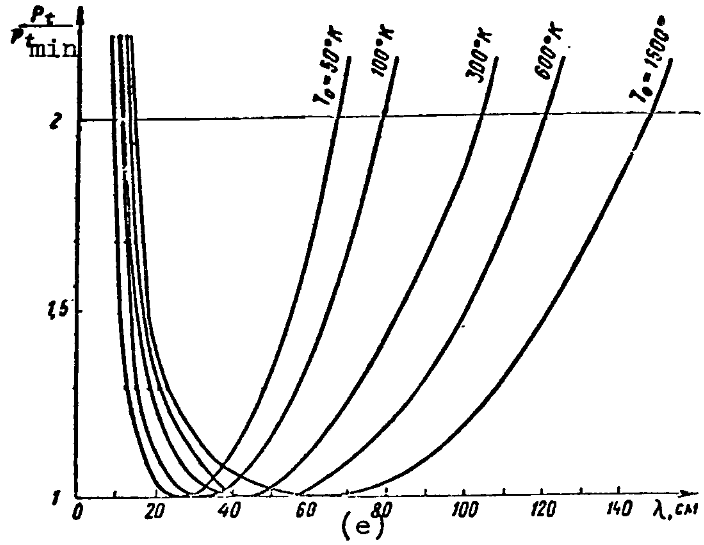
$\frac{P_t}{P_{t_{min}}}$
$T_0=50°K$
100°K
300°K
600°K
$T_0=1500°K$
2
1.5
1
0
20
40
60
80
100
120
140
λ, см

(c)



$\frac{P_t}{P_{t_{min}}}$
$T_0=50°K$
100°K
300°K
600°K
$T_0=1500°K$
2
1.5
1
0
20
40
60
80
100
120
140
λ, см

(d)



(e)

TABLE I-22. CIRCULAR EQUATORIAL ORBIT WITH h = 2,000 km.

Mode of operation with T_0, °K	FM television		PTM telegraphy		TM telephony	
	λ, cm	Δλ, cm	λ, cm	Δλ, cm	λ, cm	Δλ, cm
50	15	6-47	15	6-47	30	9-65
100	15	6-55	15	6-55	35	10-77
300	15	6-76	15	6-76	42	11-104
600	16	6-94	16	6-94	50	12-125
1,500	17	6-122	17	6-122	60	13-160

TABLE I-23. CIRCULAR POLAR ORBIT WITH h = 4,800 km.

Mode of operation with T_0, °K	FM television		PTM telegraphy		TM telephony	
	λ, cm	Δλ, cm	λ, cm	Δλ, cm	λ, cm	Δλ, cm
50	15	6-47	15	6-47	30	9-70
100	15	6-55	15	6-55	35	10-85
300	15	6-76	15	6-76	42	11-110
600	15	6-94	15	6-94	50	12-133
1,500	15	6-122	15	6-122	60	13-172

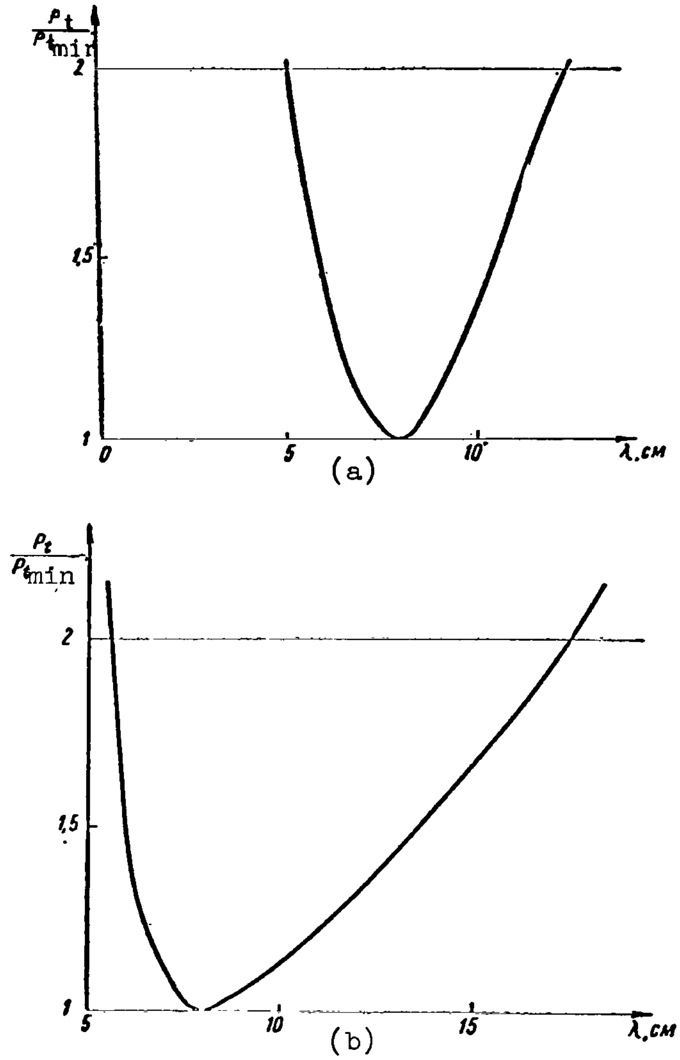
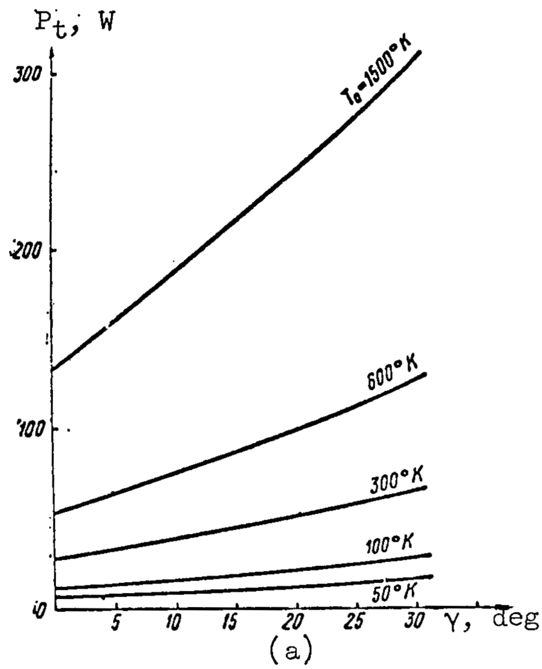
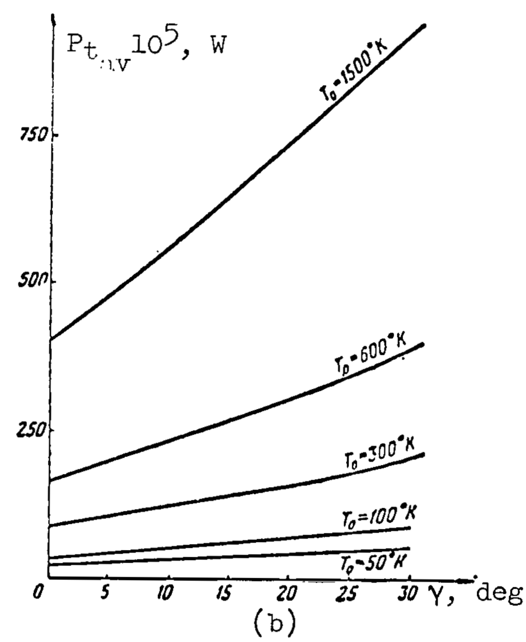


Figure I-44. Dependence of $P_t/P_{t_{min}}$ on wavelength with passive relay for all altitudes, $\nu_{des} = 10^{-4}$. a, FM television (multichannel telephony); b, FM telephony.

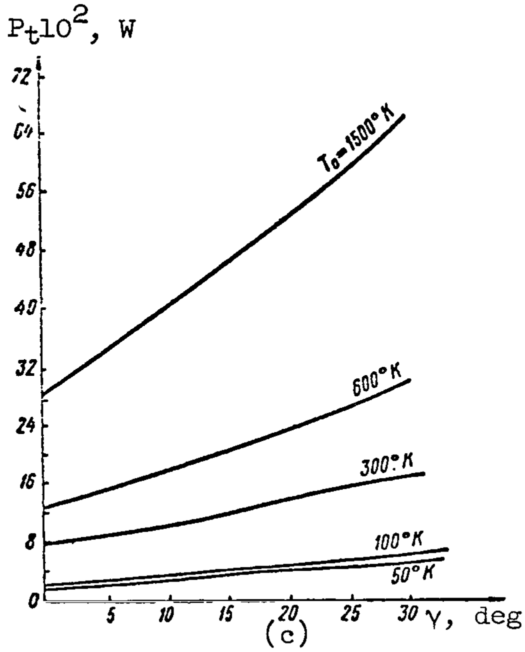


a, FM television (or multichannel SSB-FM telephony), h = 2,000 km.

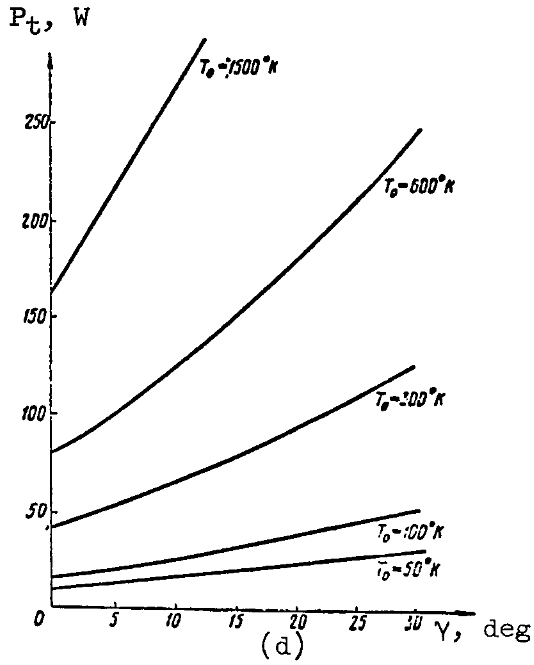


b, PTM telegraphy, h = 2,000 km.

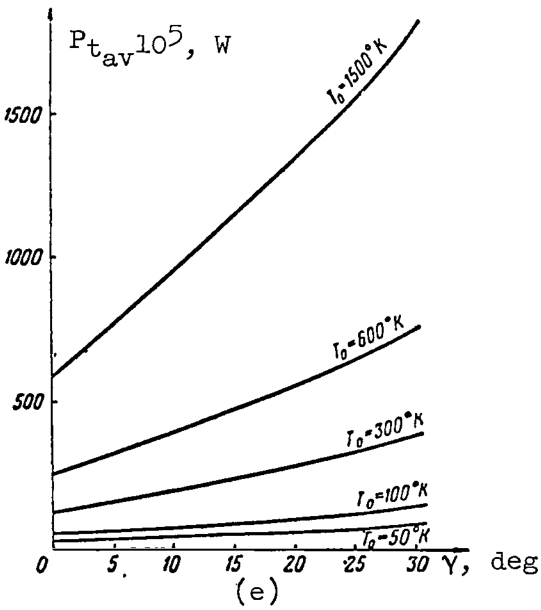
Figure I-45. Variation of repeater onboard transmitter minimal required power with accuracy of satellite pointing at Earth γ (in case of PTM average power $P_{t_{av}}$ is used) for various T_0 and orbital altitudes h.



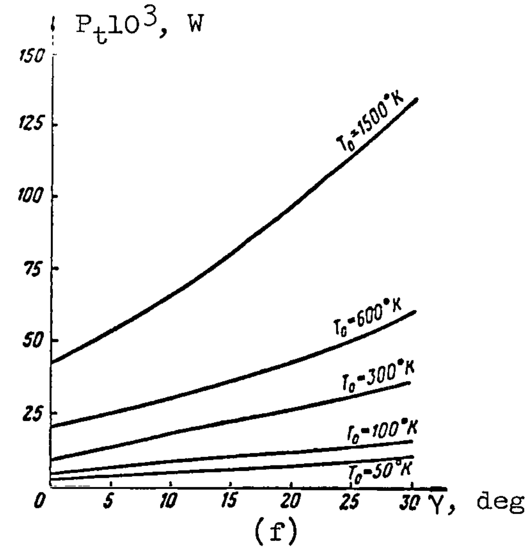
c, FM telephony, h = 2,000 km.



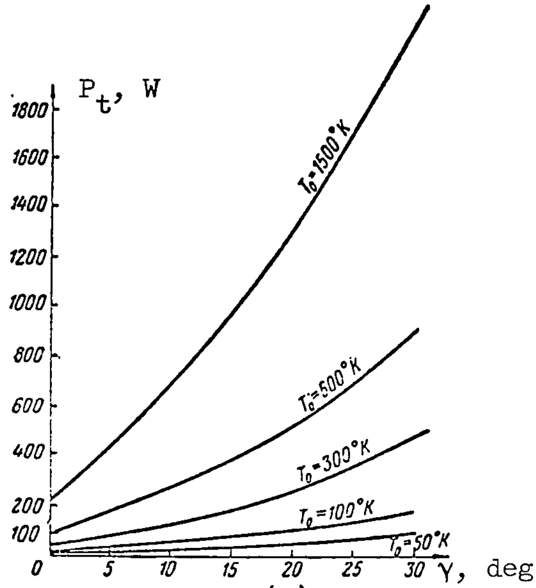
d, FM television (or multichannel SSB-FM telephony), h = 4,800 km.



e, PTM telegraphy, h = 4,800 km.

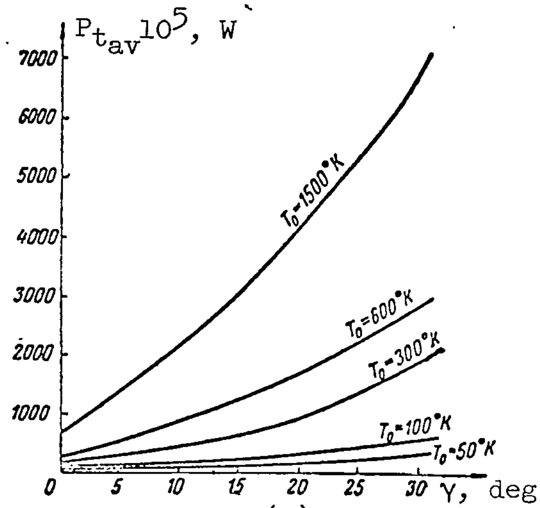


f, FM telephony, h = 4,800 km.



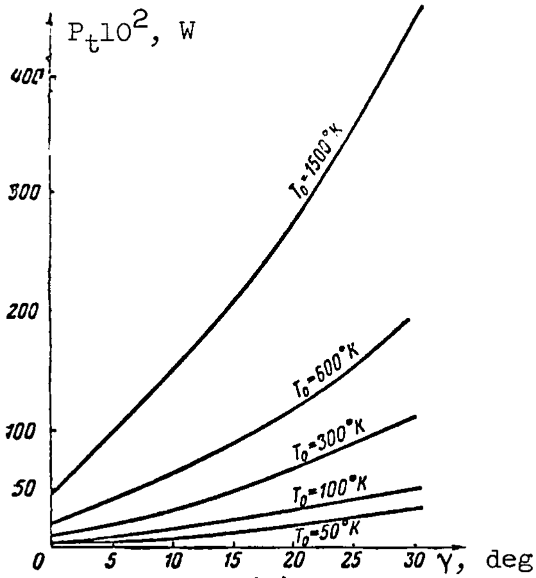
(g)

g, FM television (or multichannel SSB-FM telephony), apogee altitude h_a = 20,000 km.



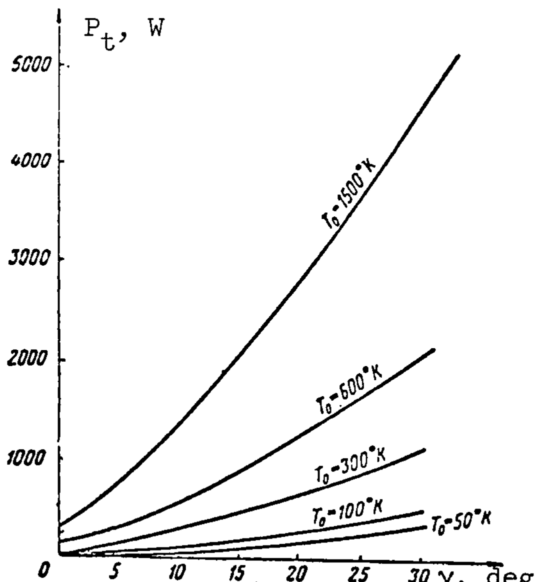
(h)

h, PTM telegraphy, apogee altitude h_a = 20,000 km.



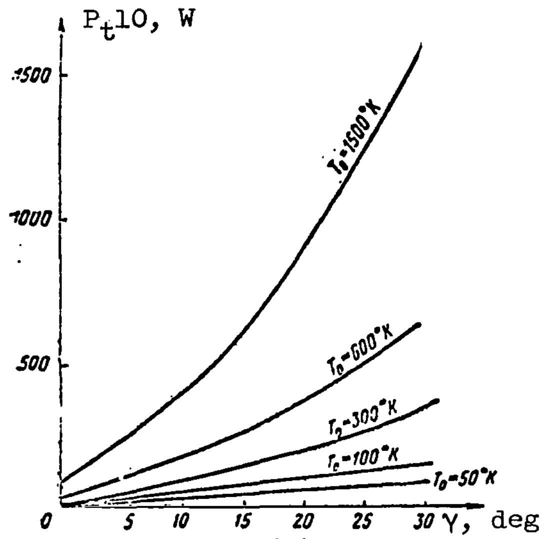
(i)

i, FM telephony, apogee altitude h_a = 20,000 km.



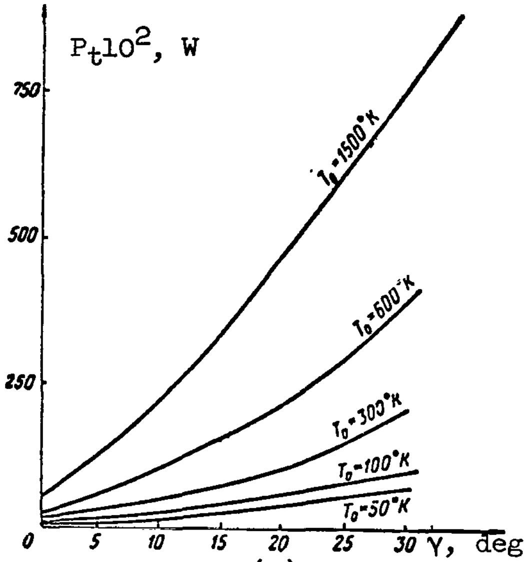
(j)

j, FM television (or multichannel SSB-FM telephony), h = 36,000 km.



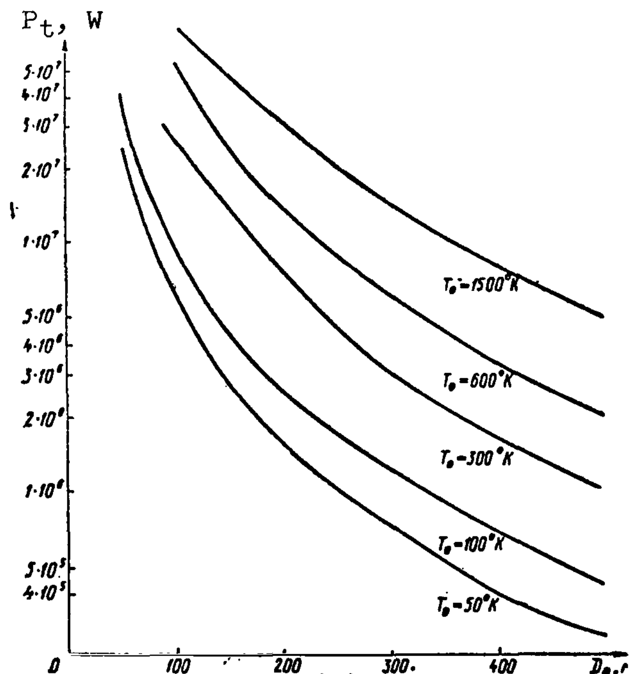
(k)

k, PTM telegraphy, h = 36,000 km.



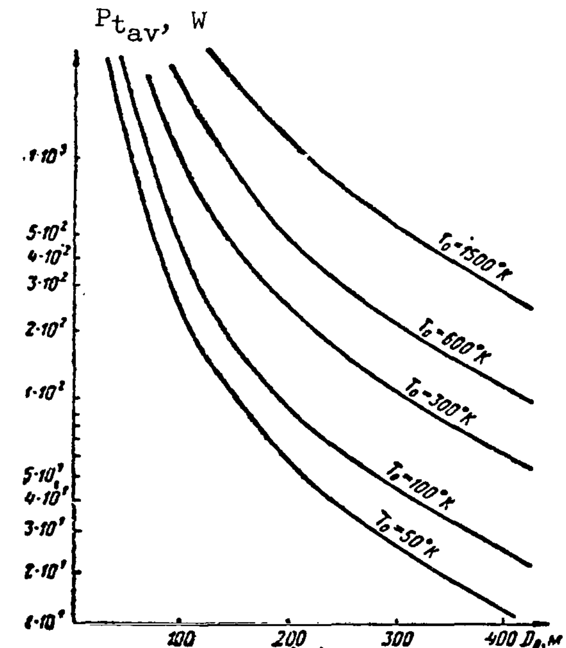
(l)

l, FM telephony, h = 36,000 km.



(a)

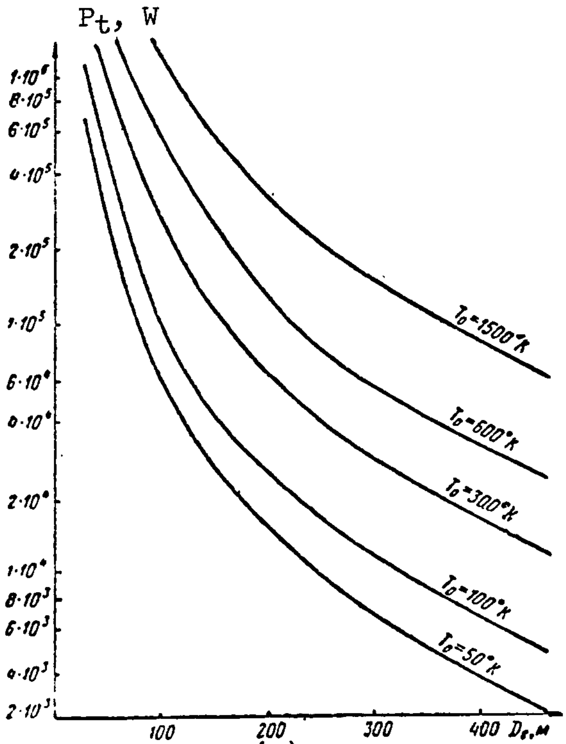
a, FM television, h = 2,000 km.



(b)

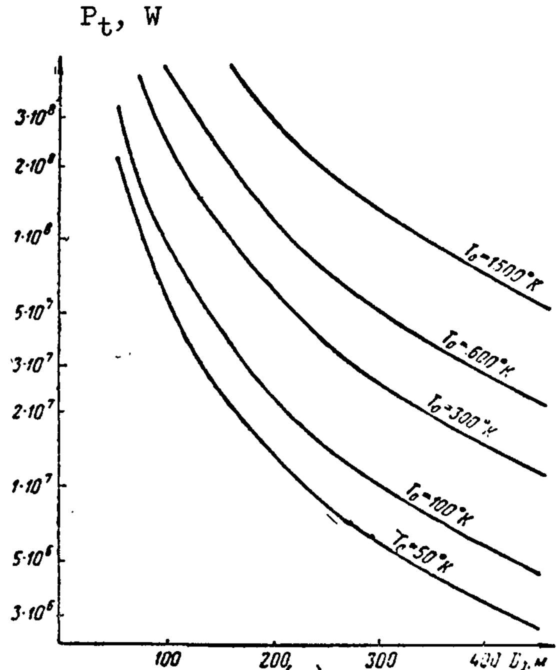
b, PTM telegraphy, h = 2,000 km.

Figure I-46. Variation of ground transmitter minimal required power as a function of diameter of passive repeater D_2 (sphere) for different T_0 and orbital altitudes h (in the case of PTM average power $P_{t_{av}}$ is used).



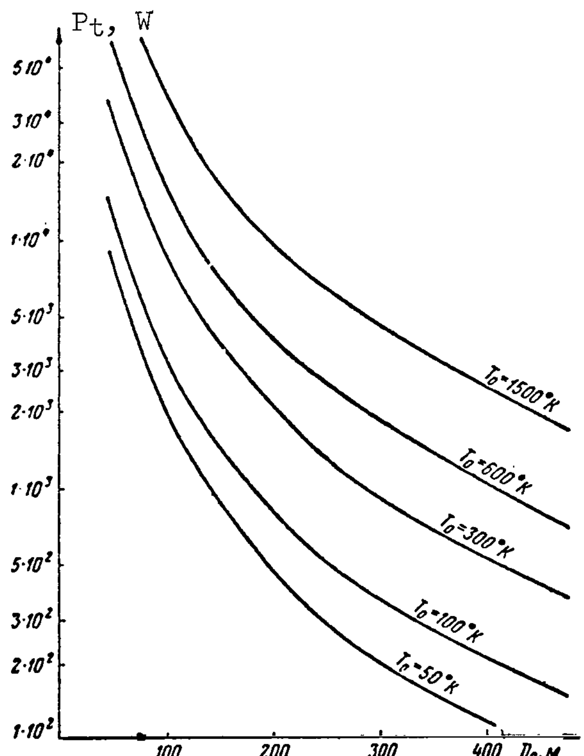
(c)

c, FM telephony, h = 2,000 km.



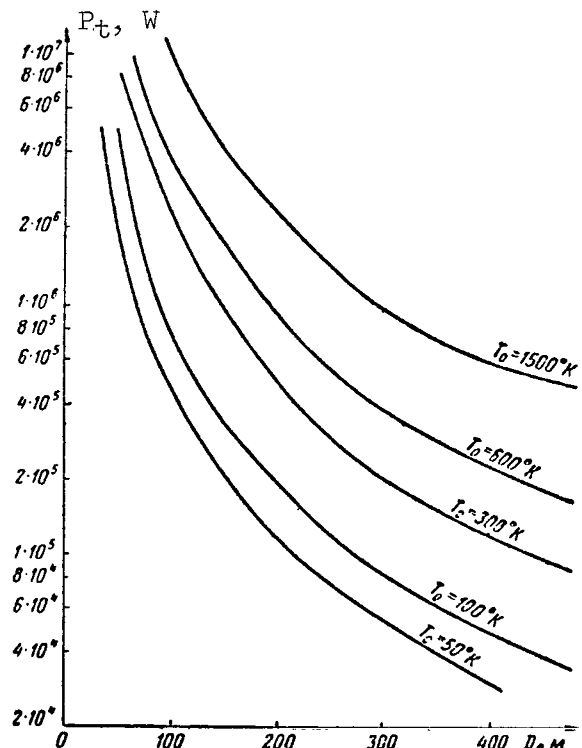
(d)

d, FM television, h = 4,800 km.



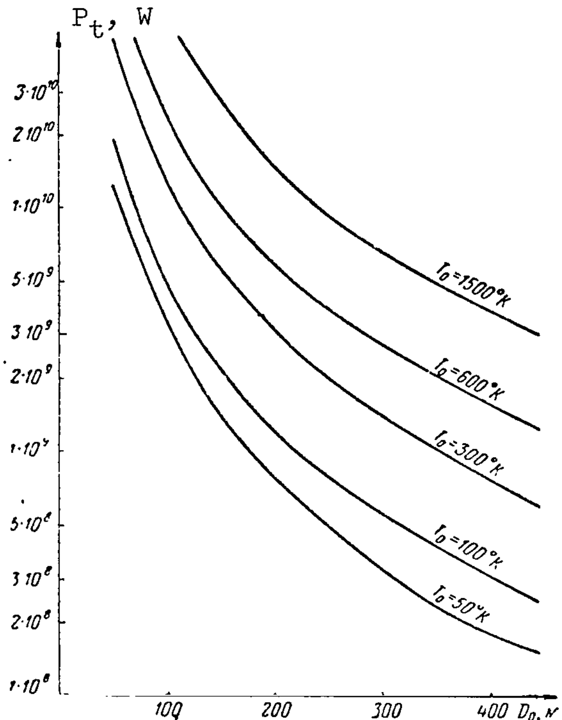
(e)

e, PTM telegraphy, h = 4,800 km.



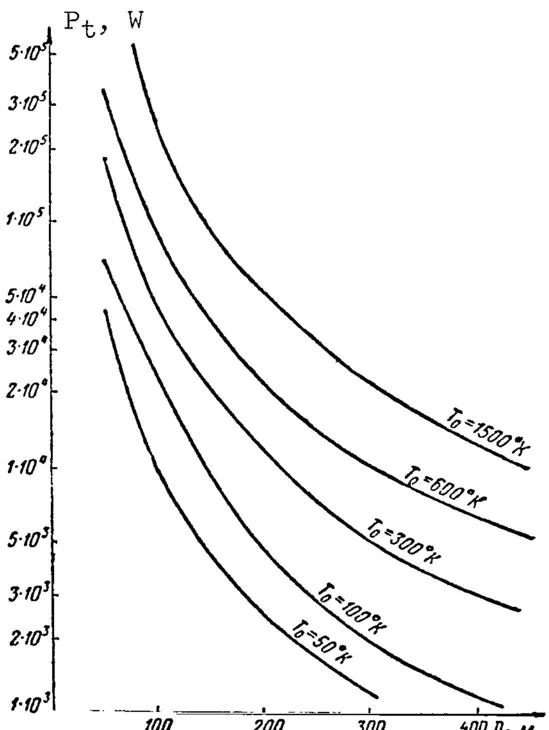
(f)

f, FM telephony, h = 4,800 km.



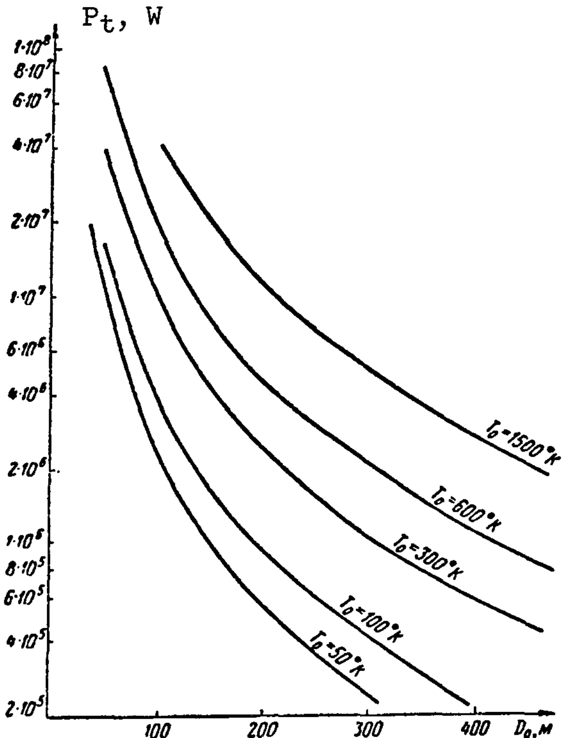
(g)

g, FM television, apogee altitude h_a = 20,000 km.



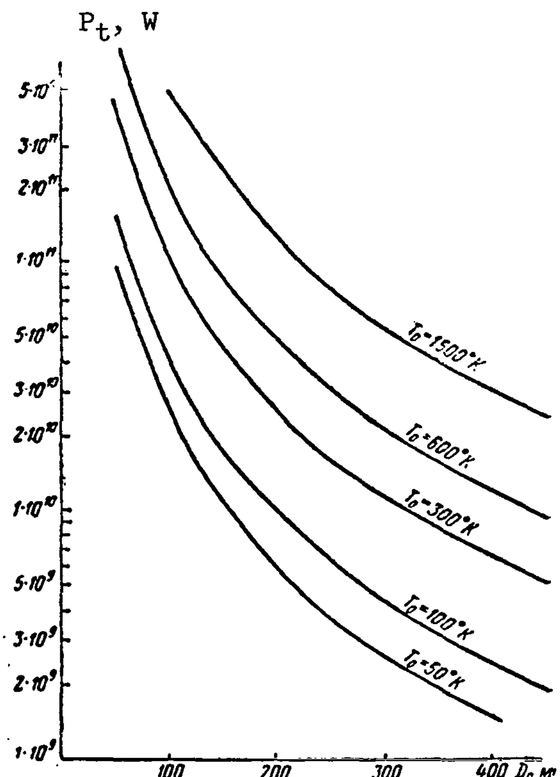
(h)

h, PTM telegraphy, apogee altitude h_a = 20,000 km.



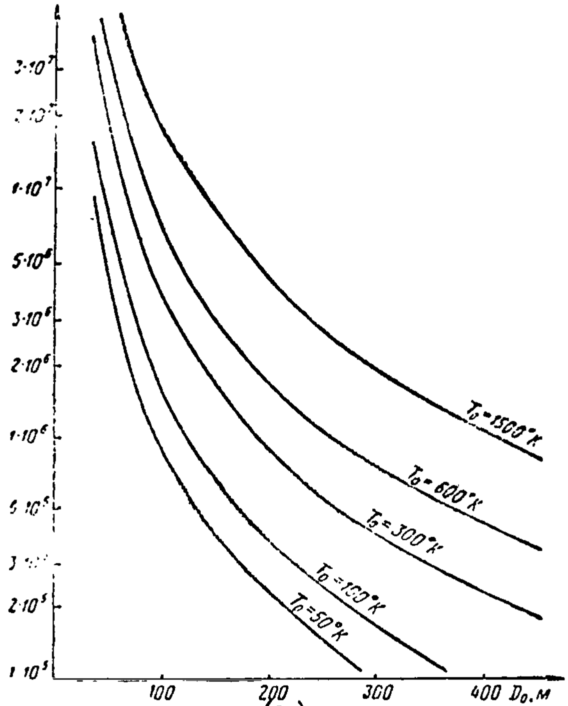
(i)

i, FM telephony, apogee altitude h_a = 20,000 km.



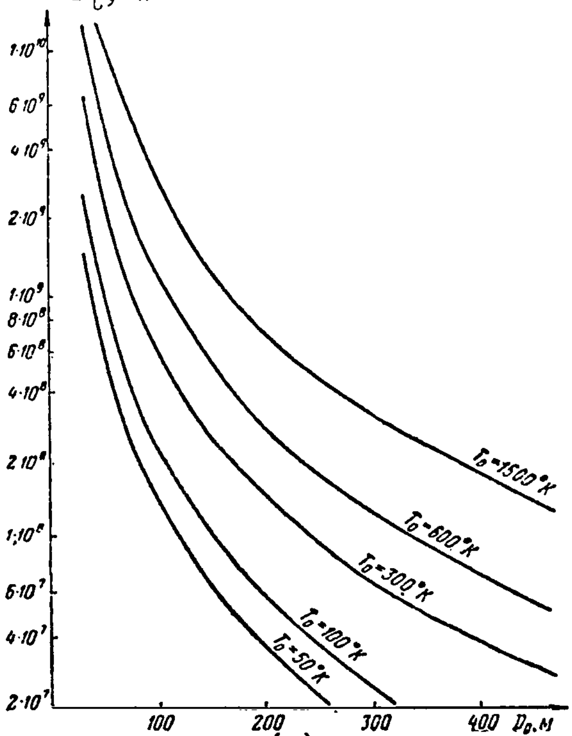
(j)

j, FM television, h = 36,000 km.



(k)

k, PTM telegraphy, h = 36,000 km.



(l)

l, FM telephony, h = 36,000 km.

TABLE I-24. ELLIPTIC ORBIT INCLINED AT 65° WITH h_a = 20,000 km.

Mode of operation with T_0, °K	FM television		PTM telegraphy		TM telephony	
	λ, cm	Δλ, cm	λ, cm	Δλ, cm	λ, cm	Δλ, cm
50	15	6-47	15	6-47	27	10-67
100	15	6-55	15	6-55	35	11-80
300	15	6-76	15	6-76	41	11-95
600	16	6-94	16	6-94	49	12-135
1,500	17	6-122	17	6-122	58	13-167

TABLE I-25. STATIONARY ORBIT WITH h = 36,000 km.

Mode of operation with T_0, °K	FM television		PTM telegraphy		TM telephony	
	λ, cm	Δλ, cm	λ, cm	Δλ, cm	λ, cm	Δλ, cm
50	15	6-47	15	6-47	25	8-70
100	15	6-55	15	6-55	35	10-81
300	15	6-94	15	6-76	42	11-108
600	16	6-94	16	6-94	48	11-125
1,500	17	6-122	17	6-122	55	12-153

6. Possibilities of Using AES as Active and Passive Repeaters

The values of the required radiation power P_t, obtained as a result of the computation, still do not give a complete picture of whether it is advisable or not, possible or not, to use, at the present stage of the development of rocket technology and electronics, artificial Earth satellites as relay stations for ground communication systems. The complete picture can be /165 obtained only by consideration of the entire problem as a whole.

We also must include in the consideration questions such as the comparison of the economic indices of the conventional long-range communication systems using UHF (radio relay links, cable lines, etc.) and the systems under discussion using relaying via AES, the possibilities of building equipments capable of operating for long periods without human intervention under the conditions of cosmic space (in the case of active relay), the possibilities of SHF generating devices (particularly in the case of passive relay), the possibilities of extended-life power sources for the onboard equipment, the possibilities of the realization of reliable systems for satellite orientation in space, etc.

The majority of these questions have received inadequate coverage in the literature and apparently require additional study.

In the examples presented above, we obtained the values of the minimal required satellite transmitter radiation power (active relay) and ground transmitter power (passive relay) for various combinations of parameters of the radio link (various modes of operation, differing values of the sensitivity of the ground receiver) and various orbits. For the radiation of this power we require the corresponding SHF generating devices and sources of power for them.

It is obvious that if none of the SHF generating devices known at the present time can supply the required value of P_t, we can immediately conclude that for the given combination of parameters of the radio link the relaying of signals via AES is not possible. Such a situation can most frequently be observed with passive relaying, since in this case the power consumption is particularly large. Therefore, we shall make an evaluation of the possibility of the use of AES as passive radio repeaters, particularly from the point of view of the capabilities of the SHF generating devices.

The relaying of signals via AES is also not possible, if the power of the source of supply for the radio equipment is not adequate to provide the required radiation power P_t. This situation is most frequently encountered /166 with active relaying, since under the onboard conditions it is difficult to provide a high-power supply with a long service life. Therefore, the evaluation of the capabilities of the use of AES as active radio relay stations is carried out from the point of view of the possibilities of building long-life power supplies for the onboard equipment.

It is evident that it is not advisable to use AES as radio repeaters in ground communication systems, if the economic expenditures on the creation of the communication system using AES are significantly higher than on the implementation of the conventional ground communication systems (cables, radio relay lines) and the technical performance of the two systems is comparable. Therefore, we shall now also consider certain economic aspects of the communication systems using AES.

Active Relay. The most important factor defining the power of the onboard transmitter and, consequently, the power potential of the space radio communication link, is the source of supply for the onboard equipment. As the result of the specific conditions of space operation, special demands are made on the power supplies for the onboard equipment.

(1) small size;
(2) lightweight;
(3) long lifetime (on the order of a year).

The onboard power supplies used up till now--batteries and conventional electric generators--do not satisfy these requirements. In this connection, intensive search is being made for methods, which are new in principle, for the conversion of various forms of energy into electrical power; work is also being done on the improvement of the existing types of sources of electric energy (refs. 73, 125 and 123).

The principal primary sources of energy for autonomous power sources are chemical, atomic and solar energy. These forms of energy can be converted to electrical energy, either directly or by means of preliminary conversion into heat, which is then converted to electric power. Figure I-47 shows the basic methods of conversion of chemical, atomic and solar energy into electric. /167

The most widely used converts of chemical energy into electrical--galvanic batteries and storage batteries--although they do continue to find wide application in rocket equipment, guided missiles and satellites as the result of the high specific power (ratio of power delivered to unit weight) and the capability of providing high current pulses for short periods of time, do have the serious drawback that they have low specific energy (energy supply per unit weight) and therefore with increase of the duration of usage their weight goes up sharply (refs. 67 and 125).

The most advanced of this type at the present time are the silver-zinc storage batteries.

Recent years have seen the intensive development of a new type of converter of chemical energy into electrical--the chemical fuel cell. In this case we first have the combustion of hydrogen in oxygen or the combustion /168 of industrial fuel gas in an air atmosphere; the heat released in these reactions is used in various ways. In some devices a difference of potentials appears directly during the combustion process. In other devices the heat released as the result of the chemical reaction is converted into electric energy by means of the Seebeck thermoelectric effect or by the thermionic emission effect.

The basic advantage of the chemical fuel elements is that they have considerably higher specific energy than the batteries and accumulators.

Since both atomic and solar energy, just as chemical energy, can also easily be converted into heat, attention is being devoted to the creation of new means of converting thermal energy into electric energy and to the improvement of the existing converters.

The conversion of heat into electricity can be accomplished with the aid of dynamic systems, i.e., the usual turbogenerators and the developmental magnetohydrodynamic generators, based on the use of the properties of plasma, or with the aid of static systems, based on the use of thermoelectric, thermionic and ferroelectric phenomena.

The electric current in the magnetohydrodynamic generator arises as the result of the energy of the moving gaseous plasma, heated to high temperature (several thousand degrees), which in its movement cuts the lines of force of a magnetic field (ref. 68). This process is similar to that of the induction of an emf in a conductor moving in a magnetic field. In the case of the magnetohydrodynamic generator, the role of the conductor is played by the ionized gas and, consequently, there are no moving parts in this generator. As a result, this generator will have outstanding reliability. On the basis of theoretical calculations, its efficiency is 40-50 percent (ref. 69).

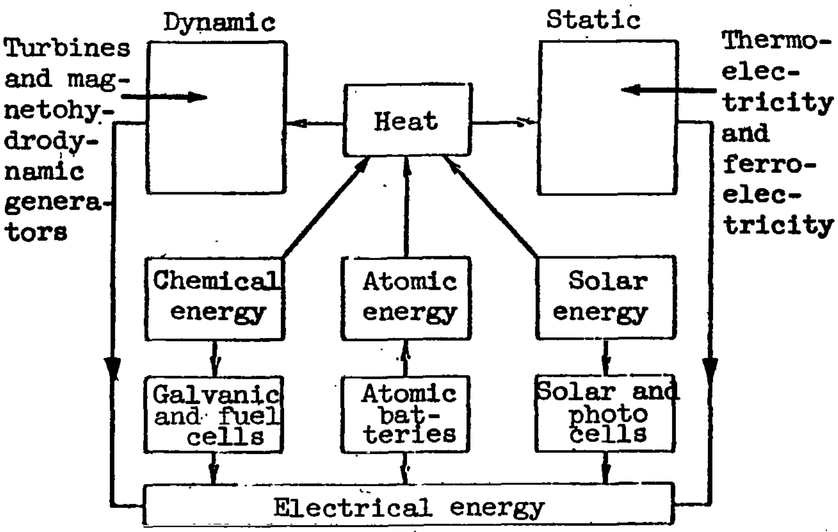


Figure I-47. Methods of converting various forms of energy into electrical energy.

Preliminary studies have also shown that the systems utilizing the magnetohydrodynamic principles can be 25 percent more efficient than the existing high-power thermal electric stations and, at the same time, considerably simpler in construction. The highest power obtained to date is 12 kW (ref. 70). /169 The thermoelectric generator consists of a series of elements, each of which is fabricated from a pair of dissimilar semiconductor materials. One junction between the dissimilar elements is heated, while the other is cooled. The electricity is generated directly as the result of the temperature difference between the hot and cold junctions in accordance with the Seebeck effect. This effect has been known for a long time, but generators based on this principle have not been used because of the low efficiency (about 1 percent). The situation changed with the appearance of the semiconductor elements. Computations show that the theoretical efficiency of the thermoelectric generators using semiconductors can reach 35 percent. According to reference 69, firms in the U.S. have already developed thermoelectric generators with an efficiency of about 10 percent with differing power output (the highest power of 500 W is that of the Westinghouse ceramic thermoelectric generator, the efficiency is 10 percent).

The generators based on the principle of thermionic emission generate electric current directly by means of the emission of electrons from a heated cathode, which are then collected by the cooled cathode. The operation of the thermionic converter is similar to the operation of the vacuum tube diode used in electronics, with the exception that no voltage is applied to the electrodes. The theoretical efficiency of the thermionic devices is 65 percent (ref. 71).

Of particular interest is the application in space of the thermionic converters with atomic reactors. One of the possible applications is the location of the ionic energy converter inside of the reactor. In this case the surface of each reactor fuel element will serve as the surface of a thermionic converter emitting electrons, and the anode will be cooled by the reactor coolant (ref. 72).

The following data give some idea of the operating characteristics of the thermionic converter: (1) the minimal temperature of the cathode which can be used for the gas-filled or vacuum converters is 1,700°K and 1,300°K, respectively; /170 (2) to minimize the specific area of the radiator surface the anode temperature must be 75 percent of the cathode temperature; (3) the converter power presently obtainable in the laboratory is 1-10 w/cm^2 of cathode area with an efficiency of 5-10 percent. Thus, with a cathode area of 1 m^2 it is possible to get a power of the order of 100 kW.

The U.S. firms RCA and Thiokol have developed an experimental version of a thermionic converter designed for installation in the exhaust duct of a jet engine operating with solid fuel. The converter consists of two coaxial cylinders 300 mm long and 44 mm in diameter. The cathode is the inner molybdenum cylinder, equipped with a tantalum sleeve to increase the emission, and heated by the exhaust gases passing through it. The anode is the outer copper cylinder, covered with nickel. The distance between the cylinders is about 1 mm. This space is filled with cesium vapors, which neutralize the space charge by ionization of the cesium atoms with impact on the surface of the heated cathode.

The conversion coefficient of the versions developed is 8 percent at a temperature of 2,473°K. The power developed is 270 W, the weight is 1.5 kg (ref. 72).

The primary advantages of the thermionic converter are the absence of large rotating mechanisms and the possibility of removal of the waste heat directly from the surface of the anode.

The generators using chemical fuel elements are electrochemical devices in which the chemical energy is converted directly into electric energy. The operation of the fuel element is based on an effect which is the reverse of electrolysis of water. In electrolysis of water, when the electric current passes through the water, the water molecules dissociate into hydrogen and oxygen; in the reverse process, when hydrogen and oxygen are combined, a flux of electrically charged particles occurs in the electrolyte, and these particles form the electric current in the external circuit. The theoretical efficiency of such elements reaches 60-70 percent. The combining of individual elements into a battery makes it possible to obtain a value of the voltage which is acceptable for operational use (24 V, for example) and adequately /171 high power, to 10 kW (ref. 69).

However, both atomic and solar energy can also be converted directly into electric energy without the intermediate stage of conversion to thermal energy.

Several methods have been proposed for the direct conversion of atomic energy into electrical energy with the use of the energy of radioactive emission of the isotopes, based on the use of:

(1) direct charging;
(2) ionization of a gas;
(3) charge carriers in semiconductor junctions;
(4) intermediate conversion of the energy of the radioactive radiation into light energy.

The conversion of the energy of the radioactive radiation of the isotopes by means of direct charging is possible only with the use of α- and β-radiation, ions and other charged particles.

In the nuclear batteries of this type the electrons, emerging with high velocity from the radioactive isotope, penetrate the insulating space and are gathered on the external conducting electrode, which becomes negatively charged. The battery has the disadvantage that it provides only small currents, since the devices based on the direct charge method have high internal resistance; this disadvantage can be overcome only by use of large quantities of radioactive material, which is hardly possible in view of its high cost.

In batteries utilizing the contact potential difference, the radioactive radiation ionizes a gas contained in the space between two dissimilar metallic electrodes. The ions and electrons formed are directed to the oppositely charged electrodes, connected through the external circuit.

The atomic batteries based on this principle also give small currents, and therefore the possibilities of their application are the same as for the direct-charge batteries.

In batteries using semiconductor junctions, the radioactive source irradiates a p-n junction made from silicon or germanium. Under the influence of the radiation, charge carriers appear in the semiconductor junction; these carriers are then separated by the field of the semiconductor junction. /172

In the batteries based on the use of the semiconductors, use can be made of radiation in the form of the charged particles (β-particles) and also the γ-radiation. With the use of silicon, we can obtain a voltage as high as half a volt at the output. The highest efficiency achievable is on the order of 3 percent (ref. 71).

The intermediate conversion of the energy of the radioactive isotopes into light energy using photocontact devices makes it possible to avoid the direct action of the radioactive emission on the semiconductor junction and prevent damage to the junction.

Table I-26 presents some characteristics of the atomic batteries of all the types mentioned above, which have been developed by various U.S. firms (ref. 71).

Analysis of the table data shows that the power output of the batteries is exceptionally low (about 1 μW). The cost of the radioactive materials is high. From the estimate of the U.S. specialists, 1 kW-hr of power from atomic batteries costs about 14,000 dollars, while the cost of using conventional /173

TABLE I-26

Battery characteristic	Principle of operation of battery					
	Direct charge			Contact potential difference	Semiconductor contact	Photo-effect
Radioactive material	Strontium-90	Strontium-90, hydrogen-3	Strontium-90, krypton-85	Hydrogen-3	Strontium-90	Promethium
Half-life, years	25	12	10	12	25	2.6
Radioactivity, curies	$10 \cdot 10^{-3}$	1	1	1.5 per cell	$5 \cdot 10^{-3}$	4.5
Size, cm^3	16.4	16.4	2	16.4	-	17.8
Weight, g	187	30	435	16.6	-	18.6
Current A	10^{-12}	$16 \cdot 10^{-10}$	10^{-9}	10^{-10}	$5 \cdot 10^{-6}$	-
Voltage, kV	14	1	1	0.1	0.2 V	0.25-1 V

methods is about 1 cent per kW-hr. In addition, there are considerable difficulties in shielding the instruments themselves from the radiation (the semiconductors, in particular). Therefore many U.S. specialists consider that batteries based on the direct conversion of atomic energy into electrical energy are not promising for use as power supplies for the onboard equipment.

The most widely used method of conversion of solar energy into electrical is the photoelectric method, based on the use of semiconductor p-n and n-p junctions in the so-called solar cells. The cell consists of two silicon layers of p and n type placed one over the other and forming a p-n junction, in which the conversion of radiant energy into electrical energy takes place (ref. 73).

The theoretical efficiency of the silicon solar elements is about 23 percent. However, a maximal figure of 15 percent is achieved at the present time (Hoffman Electronic Corp., ref. 74).

The typical silicon solar cells mass produced in 1959 had the following parameters (ref. 75):

(1) cell size 0.5 X 1 cm, 0.5 X 2 cm, and 1 X 2 cm;

(2) voltage developed 350-400 mV;

(3) eff. 8-10 percent;

(4) specific power 90 W/kg.

The power which can be taken from a panel of solar cells with a given area S depends on the magnitude of the radiant energy flux of the Sun incident on the given panel of cells. The Sun radiates 1,000 W per 1 m^2 at the surface of Earth. Outside Earth's atmosphere in the direct vicinity of Earth, the intensity of the Sun's radiation reaches a value of 1,400 W/m^2.

Thus, at the surface of Earth a solar battery cell with an efficiency of 10 percent will yield about 10 mW/cm^2. With a panel area $S = 1\ m^2 = 10^4\ cm^2$, a power $P_0 = 100$ W can be obtained.

Studies are being made of the possibilities of increasing the quantity /174 of solar energy per unit area by the use of mirrors--solar energy concentrators. This increases the cell temperature. However, with increase of the temperature, the efficiency of the solar cells decreases by about 0.6 percent for each degree centigrade (ref. 76). In spite of this, the efficiency of the entire installation can be greater than with the conventional panels of solar cells.

One of the serious problems associated with the solar elements which must be solved is that of destruction of the cells by radiation. At the present time scientists are analyzing new technology for forming the junction, in order to find a way to insure improved resistance of the solar cells to high radiation dosages.

The solar batteries have found wide application on board AES, where they are frequently used as the primary power supply sources for the electronic equipment. In these cases the storage batteries installed on board the satellites serve only as accumulators of the electric energy developed by the solar batteries. The combined use of solar batteries and storage batteries provides power to the equipment during passage of the satellite through Earth's shadow or to use the storage batteries in a pulse mode, obtaining high power from them for short periods of time.

An example of such usage is the U.S. Pioneer V solar satellite, on whose four specially designed paddles were mounted 4,800 silicon solar cells intended for supply of power to the 5 and 150 W transmitters (ref. 77).

In spite of certain advantages of the solar batteries, they cannot be considered the most promising power source for the onboard equipment in all cases. Actually, at quite large distances from the Sun the flux of radiant energy of the Sun is small, and the power obtainable from the solar batteries will not be sufficient for supplying the transmitter (all the more so, since the distance from Earth will be great and the power requirements will increase).

The consideration given here to the methods of obtaining electric power permits us to draw the conclusions: /175

1. Most promising for space conditions are the power sources in which nuclear energy is the primary source. Here it is advisable first to convert the nuclear energy into thermal energy, and then, using either thermionic, thermoelectric or magnetohydrodynamic energy converters, to convert the thermal energy into electrical (refs. 125, 123 and 122).

Of course, we must keep in mind that this form of power supply, too, has certain serious drawbacks, which hinder its application on space vehicles. In particular, a major deficiency is the fact that the nuclear processes are continuous; the nuclear battery cannot be turned off. In addition, the nuclear power sources emit corpuscular and thermal radiation requiring reliable shielding. The removal of waste heat under conditions of space flight may turn out to be another difficult problem.

2. At the present stage of development the only long-life power source for the spacecraft onboard equipment are the solar cells in combination with storage batteries (ref. 125).

Therefore, in the analysis of the possibilities of the use of AES as active repeater stations in ground communication systems, from the point of view of the capabilities of the long-life power sources for the onboard equipment, we shall assume that solar cells in combination with storage batteries are used as the power source (refs. 122, 123, 125, 73 and 103).

The possibility of use of the AES as an active repeater station from the point of view of the capabilities of the power supply for the onboard equipment (solar cells) is determined by comparing the dc power P_0 required for the given combination of parameters of the radio link with the power P'_0 on board: with a given combination of parameters of the radio link (receiver sensitivity, diameter of receiving antenna, modulation mode) ground-to-ground transmission via AES is possible if $P'_0 > P_0$.

If the onboard power supply source consists of batteries of solar cells with a given value of the power obtainable per unit area, the magnitude of P'_0 /176 is determined by the area S of the solar cell panel.

The maximal value of S is determined by the capabilities of booster technology and at its present state can hardly be taken greater than 5-10 m^2. With use of solar cells with conversion efficiency of 10 percent and orientation of the panel toward the Sun, this area can deliver a power of about 1,400 W (in the vicinity of Earth). In the further estimates we shall assume this value, remembering, however, that this value is only a rough guess.

In the determination of the required dc power P_0 from the given radiation power of the onboard transmitter P_t, we must determine the power consumption in the individual elements of the repeater. Thus, we must know the block diagram of the repeater. Three such versions have been discussed in the literature (ref. 15).

(a) Block diagram with signal demodulation (fig. I-48a),

where

M, LO_1 are the mixer and the local oscillator;

IFA is the IF amplifier;

D is the detector;

LFF is the LF filter;

LFA is the LF amplifier;

M is the modulator which modulates the HF oscillator LO_2.

It is indicated that this version of the block diagram is to be used in the retransmission of pulse signals, and also in those cases when it is necessary to alter the mode of modulation of the information on the downlink.

Thus, this particular version of the repeater block diagram can be used for the relay of telegraph information with pulse-time keying (single-channel operation).

(b) Block diagram without signal demodulation with IF amplifier (fig. I-48b), in which

M_1, LO_1 are the mixer and local oscillator;

IFA is the IF amplifier;

M_2, LO_2 are the second mixer and second local oscillator;

HFA is the HF amplifier.

It is indicated that this version of the relay block diagram is optimal for the relaying of signals with an overall spectrum width no greater than tens of megacycles. In this case the IF amplifier can be transistorized to reduce its weight, size and power consumption. /177

We can conclude that this type of repeater is suitable for the relaying of single-channel and multichannel FM telephony, and also single-channel relatively narrow-band FM television (with RF bandwidth occupying tens of Mc).

/178

(c) Block diagram without signal demodulation with HF amplifier (fig. I-48c) where

HFA_1, ..., HFA_n are HF amplifier stages;

O is a shift oscillator to decouple the input.

This version of the repeater block diagram should be used in the case when a repeater passband of hundreds or thousands of Mc is required. In this case a series of traveling wave tubes can be used as the HF amplifiers HFA_1, ..., HFA_n. This type of retransmitter can be used for the relay of single-channel and multichannel wide-band television.

In all these versions of the repeater block diagrams the primary power consumers are the output stages: the HF oscillator in case a, the HF amplifier in b and c (preferably traveling wave tubes). The power required by these stages can be determined, if we know the power (dc) supplied P_0 and the stage efficiency η

$$P_0 = \frac{P_t}{\eta} .$$

With operation in the 2,000 Mcps band and below, the efficiency of the oscillator and amplifier devices in use at the present time (klystrons, TWT, triodes) is about 10 percent. Assuming the margin coefficient for losses in the antenna feed circuit of the transmitter as 2, we find that to obtain the specified value of the radiation power of the onboard transmitter P_t the power

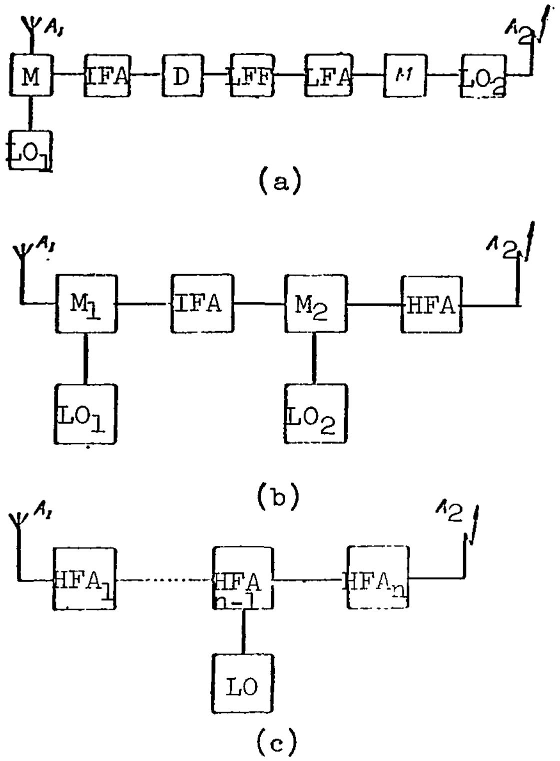


Figure I-48. a, Block diagram of repeater with signal demodulation; b, block diagram of repeater without signal demodulation with IF amplifier; c, block diagram of repeater without signal demodulation with HF amplifier.

supplied to the transmitter (dc) must be $P_0 \approx 20 \cdot P_t$ (roughly). Since $P'_{0_{max}} \approx$ 1,400 W, then $P_{t_{max}} \approx 70$ W (on the order of 50-100 W).

Let us evaluate the possibilities of the use of AES as radio relays for various cases (differing orbits, different modes of modulation), assuming $P_{t_{max}} = 70$ W.

I. Circular Equatorial Orbit with h = 2,000 km. In the case of /179
relay of FM television signals with $N_m = 10^4$, $N_0 = 16$, F = 94 Mcps, $D_2 = 20$ m, $\gamma = \pm 10^o$, $\nu_{des} = 10^{-4}$, $T_0 = 50^oK$ (no tracking filter in the ground receiver), there is a power margin of 7.4. If this margin is entirely used to reduce the

diameter of the ground antenna D_2, a ground antenna with a diameter of about 8 m is adequate, other conditions remaining the same. If a tracking filter is used in the ground receiver, there is an additional power margin of about 4. This leads to still greater simplification of the ground radio equipment--the satellite can be used for relaying of single-channel television (or multichannel telephony) with $N_0 = 2$, $N_m = 10^4$, even in the case with $D_2 = 8$ m, $T_0 = 300^\circ K$ ($\gamma = \pm 10^\circ$).

In the case of relay of FM telephony signals with $N_m = 10^4$, F = 67 kcps, $D_2 = 20$ m, $\gamma = \pm 10^\circ$, $T_0 = 50^\circ K$ (no tracking filter in the ground receiver), there is a power margin of 2,100. If the number of independent channels is 10, there remains a power margin of 210. The use of this margin permits reduction of the ground antenna to $D_2 = 1.4$ m. The use of a tracking filter simplifies the ground equipment sufficiently, so that the use of the AES for the maintenance of telephonic communication between small moving vehicles--ships, aircraft, etc., becomes possible. This conclusion is even more valid in the case of relaying of telegraphic signals.

II. Circular Polar Orbit with h = 4,800 km. In this case certain power margins are again available. They can be used just as in the preceding case.

With the relay of television signals with $N_0 = 16$, F = 94 Mcps, $D_2 = 20$ m, $\gamma = \pm 10^\circ$, $T_0 = 50^\circ K$ (no tracking filter), there is a power margin of 4.3. If this margin is entirely used for the reduction of the diameter of the ground antenna D_2, $D_2 = 10.2$ m is adequate. If a tracking filter is used in the ground receiver, the satellite can be used for the relay of single-channel /180 television (or multichannel telephony) with $N_m = 10^4$, $\gamma = \pm 10^\circ$, even in the case $D_2 = 10$ m, $T_0 = 300^\circ K$.

With the relaying of telephonic signals with $N_0 = 16$, F = 67 kcps, $D_2 = 20$ m, $\gamma = \pm 10^\circ$, $T_0 = 50^\circ K$ (no tracking filter in the ground receiver), the power margin is 1,100. If the number of independent channels is taken as 10, the remaining power margin of 110 permits reducing the diameter of the ground antenna to 2 m. With a tracking filter the ground equipment, just as in the preceding case, is simplified so that the AES can also be used for the maintenance of telephonic communication (or multichannel telegraphy) between moving vehicles. In the case of PTM telegraphy, where the power margin is still larger, the requirements on the ground receiving equipment are still further relaxed, and with a ground antenna diameter D_2 = 1 m we can use at the receiver input a crystal mixer with internal noise temperature T_0 = 1,500°K ($\gamma = \pm 10°$).

III. Elliptic Orbit Inclined at 65° with h_a = 20,000 km. In the case of relay of television signals with N_0 = 16, F = 94 Mcps, D_2 = 20 m, $\gamma = \pm 10°$, T_0 = 50°K (no tracking filter) there is practically no power margin. Reduction of the diameter of the ground antenna (reduction of cost of the ground radio equipment) can be achieved, if we increase the accuracy of the orientation of the satellite to the center of Earth to $\gamma = \pm 5°$. With $\gamma = \pm 5°$ and other conditions the same, D_2 = 12 m is sufficient.

The use of a tracking filter reduces the requirements on the ground and onboard equipment--communications are possible with $\gamma = \pm 10°$, T_0 = 300°K, D_2 = 15 m.

In the case of relay of telephonic signals with N_0 = 16, F = 67 kcps, T_0 = 50°K, $\gamma = \pm 10°$ (no tracking filter), the power margin is 670. With 10 independent channels, a ground antenna diameter D_2 = 2.5 m is sufficient, other conditions being the same. The use of a tracking filter increases the power margin, and with 10 independent channels communication is possible if /181
$\gamma = \pm 10°$, D_2 = 2 m, T_0 = 600°K, i.e., telephonic communication between moving vehicles can also be maintained.

In the case of PTM telegraphy with 10 independent channels, communication can be maintained with $\gamma = \pm 10°$, D_2 = 1.0 m, T_0 = 1,500°K.

IV. Stationary Orbit with h = 36,000 km. In the case T_0 = 50°K, D_2 = 20 m (no tracking filter) relaying of television signals (F = 94 Mcps, N_0 = 16) is possible only with an accuracy of orientation of the satellite to the center of Earth $\gamma = \pm 5°$, in contrast with the cases considered above, when an accuracy of orientation $\gamma = \pm 10°$ was adequate. The use of a tracking filter in the ground receiver increases the power margin by about a factor of 4 and communication becomes possible with $\gamma = \pm 10°$, T_0 = 300°K, D_2 = 20 m.

In the case of relaying of telephonic signals with N_0 = 16, F = 67 kcps, T_0 = 50°K, $\gamma = \pm 10°$ (no tracking filter), the power margin is 410, and with 10 independent channels it is adequate to have a ground antenna with D_2 = 3.0 m. With the use of a tracking filter communication is possible with $\gamma = \pm 10°$, D_2 = 4.0 m, T_0 = 600°K.

In the case of PTM telegraphy with 10 independent channels, communication can be maintained between correspondents with $\gamma = \pm 10^{\circ}$, $D_2 = 1.5$ m, $T_0 = 1{,}500^{\circ}$K.

This analysis permits the following conclusions. The use of AES as active relays in ground communication systems (relay of television, telephony, telegraphy) is a completely solvable problem even at today's stage of development of the technology of the construction of large ground-based antennas, sensitive receiving equipment and long-life power sources for the onboard equipment, under the condition that the satellite transmitting antenna is directed toward the center of Earth (refs. 12 and 125). Here the degree of orientation depends on the satellite flight altitude h. For relatively low satellite flight altitudes (on the order of 2,000-5,000 km) it is sufficient to stabilize the position of the vertical axis of the satellite relative to the direction to the center of Earth with an accuracy of the order of several tens of degrees. With increase of the satellite flight altitude to h = 20,000 km, the required orientation accuracy increases, γ must be on the order of 10°. In the case of the stationary satellite (h = 36,000 km) it is desirable to stabilize the position of the satellite vertical axis with an accuracy of the order of 5°, since the power requirements rise sharply with large values of γ. According to reference 78, this accuracy can be provided by gravitational stabilization without requiring expenditure of power (passive stabilization system). /182

Passive Relaying. The possibility of the use of AES as passive relay stations from the point of view of the capabilities of the microwave generating devices is determined by means of comparison of the radiation power P_t required with a given combination of parameters of the radio link with the maximal possible transmitter radiation power P_t'. This is completely defined by the capabilities of the generating device, since the capabilities of the sources of dc power are practically unlimited in this case.

Reference 79 discusses the present situation and describes the tendencies in the development of microwave amplifying and generating devices.

Analysis of the data presented in this reference shows that the klystron generators have the highest output power in the continuous and pulse regimes. The maximal values of the output power obtained today on the commercial versions of the pulsed klystrons is 20 mW at a wavelength $\lambda = 8$ cm, and 15 kW on the continuous operation klystrons (at the same wavelength).

I. Circular Equatorial Orbit with h = 2,000 km

(a) FM television with $N_0 = 16$, F = 94 Mcps, $D_{21} = D_{22} = 20$ m. Comparison of the values of P_t and P_t' shows that without a tracking filter in the receiver passive relaying of high-quality television signals via AES with a flight altitude h = 2,000 km is not possible, even with the use of high-sensitivity receiving equipment (molecular amplifier with $T_0 = 50^{\circ}$K), large /183 receiving and transmitting antennas ($D_{21} = D_{22} = 20$ m) and gigantic dimensions of the relay station ($D_0 = 400$ m).

With the use of a tracking filter there is a gain in the power of the radio link of about a factor of 4, and passive relaying of high-quality television signals with the parameters indicated above for the radio link becomes possible with four klystrons used in parallel.

(b) FM telephony with $N_0 = 16$, $F_0 = 67$ kcps, $D_{21} = D_{22} = 20$ m. Comparison of the values of P_t P'_t shows that passive relaying of high-quality telephonic signals via AES is possible even without the use of a tracking filter in the receiver, and with certain combinations of parameters of the radio link there are power margins which make it possible to relax the requirements on the ground receiving-transmitting equipment, thereby improving the economic indices of the communication system in question.

Thus, with $D_0 = 400$ m, $T_0 = 50^{\circ}$K communication is possible with the use of antennas with diameters $D_{21} = D_{22} = 18$ m at the transmitting and receiving terminals. By parallel connection of klystrons we can increase the maximal radiation power P'_t, and achieve either still more reduction of the diameter of the ground antenna or reduction of the size of the repeater.

With use of a tracking filter in the receiver, the requirements on the receiving-transmitting equipment are relaxed still further, and communication becomes possible in the case $D_0 = 400$ m, $T_0 = 50^{\circ}$K with the use of antennas with diameter $D_{21} = D_{22} = 11$ m at both ends.

(c) PTM telegraphy with $N_m = 22.5$, $F_m = 1$ kcps, $D_{22} = D_{21} = 20$ m. Passive relaying of telegraphic signals via AES with h = 2,000 km is not only possible, but there are even considerable power margins. With $D_0 = 400$ m, $T_0 =$ /184 50°K, $D_{21} = D_{22} = 20$ m and the use of a klystron with output power of 20 mW as the generator, the power margin is $1.3 \cdot 10^3$. If this margin is wholly used for the reduction of the antenna diameters, we find that telegraphic communication is possible even with $D_{21} = D_{22} = 2.0$ m.

II. Polar Circular Orbit with h = 4,800 km

(a) FM television with $N_0 = 16$, $F_0 = 94$ Mcps, $D_{21} = D_{22} = 20$ m. Comparison of the values of P_t and P'_t shows that the relaying of high-quality television signals is not possible with any realistically achievable combinations of parameters of the radio link.

(b) FM telephony with $N_0 = 16$, $F_m = 3$ kcps, $N_m = 10^4$, $D_{21} = D_{22} = 20$ m. Without a tracking filter, passive relaying of telephonic signals via AES with $h = 4{,}800$ km is possible only under the condition of the use of high-sensitivity receiving equipment and large transmitting and receiving antennas with $D_0 = 400$ m, $D_{21} = D_{22} = 20$ m, $T_0 = 50^\circ$K. The use of a tracking filter in the receiver reduces the diameters of both antennas to $D_{21} = D_{22} = 14$ m (other conditions remaining the same).

(c) PTM telegraphy with $N_m = 22.5$, $F_m = 1$ kcps, $D_{21} = D_{22} = 20$ m. Comparison of P_t and P_t' shows that passive relaying of telegraphic messages via AES with $h = 4{,}800$ km is possible, and with certain combinations of parameters of the radio link there are power margins.

Assuming that a pulsed klystron with output power of 20 mW is used as the generator and that $T_0 = 50^\circ$K, $D_0 = 400$ m, $D_{21} = D_{22} = 20$ m, we find that the power margin is $1.7 \cdot 10^2$. If this margin is fully used for the reduction of the diameters on both ends, telegraphic communication is possible with $D_{21} = D_{22} = 5.5$ m (other conditions being constant).

III. Elliptic Orbit Inclined at 65° with $h_a = 20{,}000$ km /185

(a) FM television with $N_0 = 16$, $F_0 = 94$ Mcps, $N_m = 10^4$, $D_{21} = D_{22} = 20$ m. Comparison of the values of P_t P'_t shows that in this case passive relaying of the television signals is not possible with any realistically achievable combinations of parameters of the radio link.

(b) FM telephony with $N_0 = 16$, $F = 67$ kcps. Comparison of the values of P_t and P'_t shows that without a tracking filter passive relaying of telephonic signals is possible only with large receiving and transmitting antennas and a high-sensitivity receiver. For example, with $T_0 = 50^\circ$K, $D_0 = 400$ m, the diameter of the ground antenna must be $D_{21} = D_{22} = 74$ m. With a tracking filter in the receiver the diameters of the antennas at both ends can be reduced to $D_{21} = D_{22} = 53$ m (other conditions being the same). Reduction of the ground antennas to $D_{21} = D_{22} = 30$ m can be achieved by parallel connection of 10 klystrons.

(c) PTM telegraphy with $N_m = 22.5$, $F_m = 1$ kcps. Comparison of P_t and P'_t shows that in this case passive relaying of the telegraphic signals is quite possible, and that with certain combinations of parameters of the radio link there are power margins.

If a pulsed klystron with output power of 20 mW is used as the generator and $T_0 = 50^{\circ}$K, $D_0 = 400$ m, we find that we must use antennas with diameters $D_{21} = D_{22} = 12$ m at both terminals.

IV. Stationary Orbit with h = 36,000 km

(a) FM television with $N_0 = 16$, $F_0 = 94$ Mcps, $N_m = 10^4$. Comparison of the values of P_t and P'_t shows that passive relaying of television information in this case is not possible with any realistic combinations of parameters of the radio link.

(b) FM telephony with $N_0 = 16$, $F_0 = 67$ kcps, $N_m = 10^4$. Comparison of P_t and P'_t shows that passive relaying of telephonic signals without a track- /186 ing filter is possible only with large receiving and transmitting antennas and the use of receivers with high sensitivity. For example, with $T_0 = 50^{\circ}$K, $D_0 = 400$ m, the diameter of the antennas must be $D_{21} = D_{22} = 115$ m. The use of a tracking filter reduces the power consumption of the radio link by about a factor of 4, which, with other conditions remaining the same, permits a reduction of the diameter of both antennas to 82 m. Parallel connection of 10 klystrons makes it possible to reduce D_{21} and D_{22} to 45 m.

(c) PTM with telegraphy $N_m = 22.5$, $F_m = 1$ kcps. Comparison of the values of P_t and P'_t shows that passive relaying of telegraphic signals in this case is possible, but with the use of large receiving-transmitting antennas and huge dimensions of the repeater. For example, with $D_0 = 400$ m, $T_0 = 50^{\circ}$K and use of a pulsed klystron with output power of 20 mW as the generator, it is necessary to have antennas (at both ends) with $D_{21} = D_{22} = 20$ m.

This analysis permits us to conclude that the use of AES as passive repeaters in global and local ground communication systems is a resolvable problem at the current stage of the development of the technology of the construction of large antennas and powerful generating devices only in the case of the relaying of relatively narrow-band signals (telephony, multichannel and single-channel telegraphy) and with the use of low-flying satellites (with h = 2,000 km, 4,800 km) in contrast with active relaying, where it is already possible at the present to relay broad-band signals (television, multichannel telephony) (refs. 121, 125, 115 and 116).

This circumstance is undoubtedly a major drawback of the communication systems using passive relaying. However, the communication systems using passive relaying do have several advantages, which will be discussed below.

Reduction of the power consumption in radio links with passive relaying of the signals via AES can be achieved by means of the use of reflectors of special shape, which concentrate the radiation in the solid angle from the satellite subtended by Earth. The gain is particularly large in the case /187 of high altitude orbits. For example, in the case h_a = 20,000 km the maximal gain is a factor of 66.5, in the case h = 36,000 km the maximal gain is a factor of 162. Thus, the use of reflectors of special shape, oriented toward Earth, makes it possible to use even the stationary satellites for the relaying of relatively narrow-band signals. With D_0 = 400 m, T_0 = 50°K, antennas with diameters $D_{21} = D_{22}$ = 32 m are needed for the relaying of telephonic signals using a stationary satellite. With the use of a tracking filter, $D_{21} = D_{22}$ = 20 m antennas are required.

Economic Aspects of Communication Systems Using Artificial Earth Satellites. In the course of the development of the basic technical principles of the design of communication systems using AES, along with numerous problems, there stand out the questions of the technological and economic effectiveness of these systems, the evaluation of their cost, and the selection of the optimal parameters from the point of view of the lowest costs in the construction and operation of the new communication systems.

In the analysis of these questions we must take account of all factors which influence the economics of the communication systems using AES. The system designer must make preliminary estimates of all initial data, for example, the communication mode, duration, number of channels, location of the ground equipment, climatic conditions, service life of equipment, expected transmitter power and receiver sensitivity on board the satellite. The last two factors affect directly the size and cost of the ground antenna structures, transmitters and receivers. Low power of the satellite transmitter requires large antenna structures on the ground. And the antenna cost is closely related with its dimensions, increasing in proportion to the square of the antenna diameter.

All elements of the system are closely interrelated. The cost of the ground facilities is quite dependent on the satellite repeater mode selected (active or passive). The choice of orbit and the number of satellites affects the design of the satellite itself. For example, the higher the altitude of the orbit, the larger the nominal power requirement of the satellite transmitter, consequently the higher the satellite weight. The total weight of the satellite launched into orbit determines the size and number of stages of the booster; the size of the booster may influence the launching method, etc. /188

Consequently, the selection of the economically optimal version of the design of a communications system using AES is associated with the consideration of a large number of variants of the system design, with the study of the degree of influence of a large number of factors, and is a most complex problem with many variables. Its resolution requires good technical knowledge, working out the methodological questions of the solution, finding the functional interrelation between the technical and economic parameters of the system and accumulation of the necessary statistical material from experimental operation of similar systems. Unfortunately, at the present time we lack adequately substantiated information on such important questions, for example, as the lifetime of the various satellites, how the launch cost varies as a function of the type and weight of the satellite, the probability of successful satellite launch, the actual handling capacity of communication systems using AES, etc. Therefore, in all the known studies to date on this subject consideration is usually given only to the economic indices of individual versions of the system design using AES. These are compared with one another and with the familiar ground communications facilities, and first attempts are made at economic optimization of certain components and portions of the satellite communication systems. However, even this study is of certain value, since it permits estimation of the order of magnitude of the costs of the construction of the satellite communication system, comparing this with the costs of construction of the conventional ground communication systems.

Previously, we introduced as an example the version of a global communication system using three equidistant stationary satellites. It is of interest to consider the economic aspects of the implementation of this system.

To do this we shall make use of the results of reference 80. Here a satellite communication system is considered which includes a single communications satellite in a 24-hour equatorial orbit (stationary satellite) over the Atlantic Ocean with a handling capacity of 4,800 duplex telephone channels. The system includes seven pairs of ground stations: New York - London, New York - Bonn, New York - Paris, New York - Rome, Washington - London, Chicago - London, and Miami - Rio de Janeiro. Three such satellites are required for worldwide communications, but the average cost of a single communications channel (selected by the author as a basis for comparison) will not differ significantly. Comparison of the economic indices of the communications system using stationary satellites is made with the economic indices of the underwater cables and radio-relay lines. The cost of the cable is taken on the basis of the data of the American Telephone and Telegraph Corporation on a new type of cable, which costs less by about a factor of 9 than existing types. /189

Interest on capital invested is taken as 8 percent, and the lifetime of the ground equipment is taken as 15 years. The comparison of the indicated communications systems is made on the basis of yearly costs of a duplex channel, indirectly taking account of the magnitude of the initial capital expenditures.

Figure I-49 shows the economic indices of the communications systems compared, showing yearly costs on the telephone channel with full utilization of the handling capacity of this communication system. The economic indices of the satellite communications systems are presented for varying satellite lifetimes T and for various probabilities of successful launch of the satellites P (P = 0.5 to 0.9). The economic indices of the submarine cables and the radio relay lines are given as a function of the line length.

We can see that with full utilization of the handling capacity of the communication systems even with very pessimistic assumptions relative to the average lifetime of the satellites, the probability of their successful launch and the costs, the creation of transoceanic communication systems using AES is more economical than the laying of submarine cable even over relatively short distances. In the case of the use of satellite communication systems in regions where the major portion of the territory is land, they can compete with the conventional ground communication systems (radio-relay lines in the present case) only with the most optimistic assumptions (long satellite lifetimes, nearly unity launch probability, very long radio links). /190

At the present moment it is difficult to imagine that the capacity of the communication system via AES will immediately be fully utilized. It is

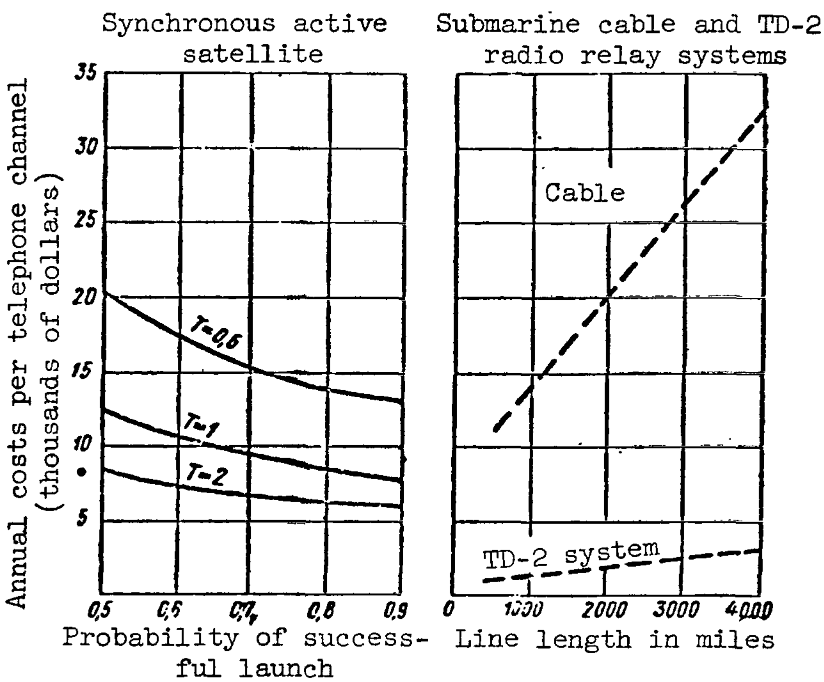


Figure I-49. Annual cost of telephone channel with full utilization of communication system capacity.

realistic to assume that the handling capacity of the satellite communication system will increase gradually. Calculations were made of the economic indices of satellite and conventional communication systems on the assumption that the initial capacity of the communication system, using a stationary satellite, will be 600 channels, and that in the course of 15 years it will increase /191 to the maximal value of 4,800 channels with an average yearly growth of 15 percent. The results of the calculations are shown in figure I-50. We see that the yearly costs per telephone channel with an average satellite lifetime of one year and a successful launch probability of 0.75 is equal to 27,500 dollars, which is considerably higher than the cost with full utilization of the capacity (10,000 dollars). Thus, the degree of utilization of the satellite communication system capacity is a very important factor in determining the economic advisability of the building of such systems.

The value of 15 percent, characterizing the rate of growth of the handling capacity of the system, is based on the data (ref. 80) of the study of U.S. specialists of the transoceanic incoming and outgoing communications in the 1930-1959 period and the estimated values for 1960-1980 (fig. I-51).

Reference 80 presents the determination of the overall costs of the creation of a global communication system using stationary satellites and the return on the investment. It is assumed that in the first stage of construction of the system it will consist of 11 large and 23 small ground stations /192 and three stationary satellites, each of which is located "fixed," respectively, over the Atlantic, Pacific and Indian Oceans. The small ground stations will be built in countries with light international traffic. They will handle about 13 percent of the total international traffic of the satellite system using the three satellites. The large ground stations are proposed for countries with heavy international traffic.

The capital investments required for the construction of the small and large ground satellite communication stations are 928,000 dollars for the small stations and 5,635,000 dollars for the large stations. The cost breakdown is shown in table I-27. /193

The yearly costs for the ground stations are respectively 301,000 and 1,520,000 dollars for the small and large stations. Here the direct expenditures on the technical operation amount to 81,000 and 226,000 dollars per year, respectively, for the small and large stations. The initial costs of the launch of the satellites amount to 78.5 million dollars. The breakdown of the costs is given in table I-28.

The annual costs for the communication satellite launches are 52.06 million dollars, on the assumption that the service life of the satellites is two years, and 41.06 million if the satellite service life is taken to be three years.

The overall costs of the construction of the satellite communication system are approximately 160 million dollars. In order to determine the economic effectiveness of the satellite communication system, calculations were made of the minimal number of channels at each ground station, which will provide for

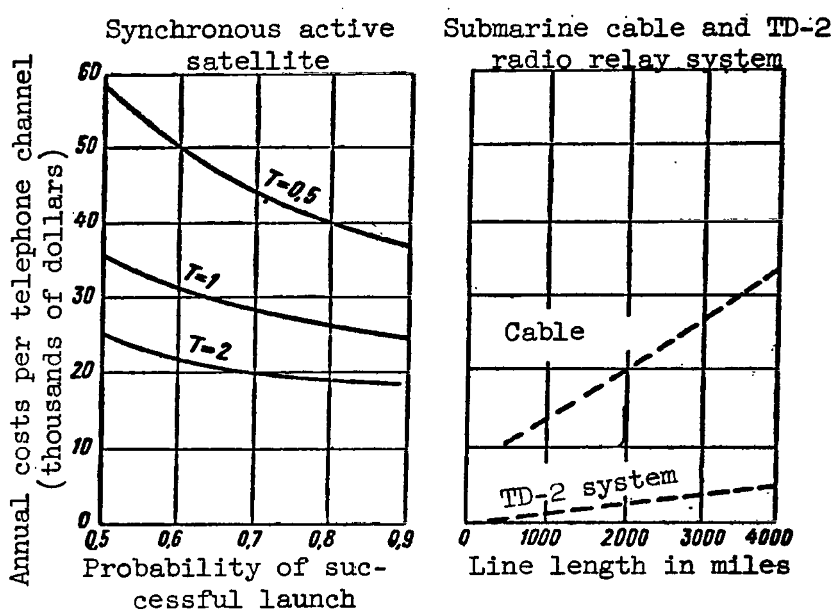


Figure I-50. Annual costs per telephone channel with gradual growth of utilization of communication system capacity.

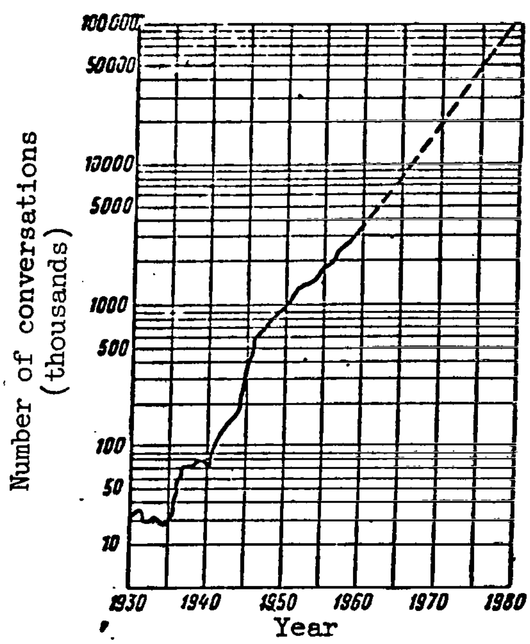


Figure I-51. U.S. transoceanic incoming and outgoing telephone traffic for 1930-1959 and estimated traffic for 1960 to 1980.

TABLE I-27

Cost item	Small station		Large station	
	Number of units	Total cost, thousands of dollars	Number of units	Total cost, thousands of dollars
Land (if not government owned)		10		50
Access road	1.6 km	20	16 km	250
Building		80		250
Power amplifiers and heat exchangers	2	100	8	1,500
SSB oscillators	2	20	8	80
Receivers with molecular amplifiers	2	40	8	160
Multiplexing equipment	2 X 12 channels	120	2 X 300 channels	1,200
Fixed antenna	one, $D_2 = 10$ m	55	one, $D_2 = 20$ m	200
Echo-suppressor	12	12	150	150
FM receiver		7		60
Audio frequency generator equipment		15		1,000
Television exciters				60
Control equipment				100
Monitor-testing equipment		50		100
Communication to nearest center	1st part	50	2nd part	100
Electric and power supply		65		135
Other costs		284		340
Total		928		5,635

TABLE I-28 /194

Cost item	Total cost, millions of dollars
Satellites and launchings	66.0
Three main control stations	10.5
Computers	1.5
Engineering design cost	0.5
Total:	78.5

the amortization of the funds invested, and will provide for an income of 15 percent of the capital invested. It was found that the minimal number of channels is 13 or 11 for the small stations, depending on the satellite lifetime (2 or 3 years). For the large stations the number is 242 or 200 channels, respectively.

Based on the requirements of the U.S. for internal and external telephone traffic and on the assumption that 25 percent of the communications will be accomplished using the satellite communications system with synchronous orbits, it was established that at each of the five large stations in the U.S. there will be needed:

in 1965 - 30 channels
in 1970 - 125 channels
in 1980 - 648 channels

From this the authors conclude that by about 1967 in the U.S. it will be possible, but not economically justified, to build five ground stations equipped with 36 multiplexed channels. By 1970 the construction of five ground stations, each equipped with 120 multiplexed telephone channels, will be /195 justified.

Reference 81 presents a description and economic indices of a global communication system using satellites in nonsynchronous orbits. The system consists of 12 satellites launched into circular equatorial orbits with a height of about 7,500 miles (12,000 km) with an orbital period of 12 hours. Each satellite weighs about 16 kg.

The region served by the satellite communication system is divided into 6 zones, whose terminal stations are in direct communication with one another. Retransmitting stations located on the boundaries of the zones provide communication between the zones. The total number of terminal stations is 70, the number of retransmitting stations is 6. At the terminal stations there are three mobile antennas of 50 ft diameter (15 m) each (two operational and one standby), and at the retransmitting stations there are four mobile antennas of 80 ft (25 m) diameter each (three operational and one standby). A small computer is installed at each ground station for the control of the antennas.

The planned communication system capacity for the early stages of construction is 600 duplex channels, increasing to 1,200 duplex channels in the later stages, plus one or more television channels. It is presumed that the information flow will primarily be closed within the zones. Interzonal traffic will amount to less than 15 percent.

On the assumption that the satellite service life is four years, the probability of successful launch is 0.75, and the service life of the ground facilities and equipment is 20 years, it is found that the total costs for the system over 8 years will be 193.4 pounds sterling.

The total number of channels in this system will be 960 in 1970 and 2,000 in 1980. Assuming that the usage of each channel is 45,000 minutes per year and that the tariff is 15 shillings per minute of conversation, we find that over 8 years of operation the income is about 370 million pounds sterling. This value is greater than the total overall costs of the construction and operation of the satellite system, which is indicative of the economic /196
effectiveness of the communication systems in question.

The consideration of certain economic aspects of the satellite communication systems makes it possible to draw the following conclusions (refs. 80 and 81).

1. With accounting for the high rates of growth of the requirements for international telephone traffic, the construction of satellite communications systems is economically justified (comparatively rapid amortization of the communication system is provided).

2. The satellite communications system is economically more favorable than the systems using submarine cable, even with comparatively short communication distances (on the order of 1,000-2,000 km).

3. The satellite communications systems are more favorable than the radio-relay systems only over long distances (on the order of 10,000 to 15,000 km), i.e., in those cases when communication is to encompass a vast region of Earth's surface.

Summarizing the discussion of the possibilities of the use of AES as relay stations in ground communication systems, we can draw the general conclusion that even at today's stage of the development of rocket technology and electronics the creation of ground communication systems (global and local) using AES is fully practicable technically and is economically advisable.

CHAPTER II. USE OF THE MOON AS A RELAY STATION FOR SURFACE COMMUNICATIONS SYSTEMS

1. General Information on the Moon

In addition to artificial Earth satellites, the natural satellite of the Earth--the Moon--can also be used as a relay station for surface communications systems. Table II-1 presents the basic data relating to the Moon (ref. 82). /197

The Moon, just as the Sun, travels across the sky continuously from west to east and describes a large circle in about one month. The displacement of the Moon amounts to 12-13° per day. We differentiate the sidereal period of revolution (or sidereal month) and the synodic month. The first is equal to the period of revolution of the Moon about the Earth, at the end of which the Moon returns to the same position among the stars (in the mean equal to 27.32 mean solar days). At the end of /198 the synodic month (in the mean equal to 29.53 mean solar days) the Moon returns to the same position in space relative to the Sun and the Earth.

Detailed study discloses complex characteristics of the lunar movement. The visible path of the Moon changes every month and its movement along this path takes place nonuniformly. The deviations from uniform motion of the Moon across the sky are termed inequalities. The inequalities are caused by the fact that the Moon does not travel in a circle, but in a first approximation along an ellipse with an eccentricity equal to 0.055, and by the disturbances from the Sun and the oblateness of the Earth.

TABLE II-1

Parameter	Value
Distance from Earth, km	
at perigee	356,400
at apogee	406,670
average	384,900
Period of revolution about Sun, days	27.3
Orbital eccentricity	0.055
Mean orbital velocity, km/sec	1.02
Equatorial diameter, km	3,473
Mass with relation to mass of Earth	0.01228
Surface gravity force with relation to gravity force on Earth's surface	0.166

Therefore, for the description of the motion of the Moon we must have recourse to the concept of the osculating ellipse. We can say that the Moon moves along an ellipse which rotates, first, in its plane from west to east (with a period of about 9 years) and second, it moves simultaneously about the ecliptic axis from east to west with a period of about 18.6 years (ref. 83).

The lunar orbit is located in a plane inclined at a small angle to the plane of the orbit of the Earth about the Sun. This angle does not remain constant and varies with a period of 18.6 years in the range from $4^{\circ}59'$ to $5^{\circ}17'$. On the average it is $5^{\circ}9'$. Thus, the Moon is practically an equatorial satellite.

We must recall one interesting peculiarity of the motion of the Moon: as it revolves about the Earth, the Moon keeps the same side turned toward the Earth at all times. This is explained by the fact that the period of revolution of the Moon coincides exactly with the period of its rotation about its axis. However, the orientation of the Moon toward the center of the Earth is not precise: thanks to the so-called libration phenomenon, a small portion of the other hemisphere of the Moon can be seen. The libration phenomenon amounts to the following. The Moon covers different portions of its elliptic orbit with differing velocity: faster at the perigee, and slower at the apogee. However, the rotation of the Moon about its axis is uniform. Therefore, in some portion of a revolution in its orbit it sometimes turns less on its axis than it turns in the portion of the revolution and sometimes more, and a small portion of the other hemisphere is turned toward the Earth. This phenomenon has been termed longitude libration. There is also /199 latitude libration, associated with the fact that the plane of the lunar orbit is inclined to the equatorial plane of the Moon by an angle of $6^{\circ}30'$. Therefore the regions of the back hemisphere which are located beyond the poles become visible. The libration phenomena make it possible to see about 9 percent of the lunar area in addition to the hemisphere turned toward the Earth.

The velocities of the lunar libration ι_t lie approximately in the range of $(5\text{-}10)10^{-7}$ rad/sec.

In considering the possibilities of the use of radio links with relay of the signals via the Moon, we can again analyze the cases of active and passive relay. We shall analyze these two cases separately.

2. Active Signal Relay Via the Moon

Assuming that an automatic radio relay station has been established on the Moon, we shall make a power requirement analysis of the Moon-Earth radio communication link as a function of the nature of the information transmitted and the parameters of the ground receiving and antenna equipments.

We should note that the case of relaying of signals via the Moon practically coincides with the case of active relaying of signals via AES.

Their difference consists only in that in the calculation of the Moon-Earth radio link we must take account of the intrinsic noise of the Moon ($T_{ex} = 220^{\circ}K$).

The presence of the additional term in the expression for the effective noise temperature at the receiver input T_{eff} leads to the optimal operational frequencies of the Moon-Earth radio link for certain combinations of radio link parameters being somewhat different from the optimal operational frequencies of links with relaying via AES.

We shall perform the analysis of the optimal wavelengths for the Moon-Earth radio link for various forms of modulation (FM television and telephony, PTM telegraphy), for a series of values of T_0 ($50^{\circ}K$, $100^{\circ}K$, $300^{\circ}K$, $600^{\circ}K$, $1,500^{\circ}K$) and $\delta_{min} = 5^{\circ}$ (minimal antenna elevation angle). We summarize the results in table II-2.

Let us analyze the required transmitter radiation power of the lunar retransmitter for the cases of relay of television (FM, $N_0 = 16$, $F_0 = 94$ Mcps), telephony (FM, $N_0 = 16$, $F_0 = 67$ kcps), PTM telegraphy (pulse-time keying, pulse duration $\tau = 1$ microsec, duty cycle 1000, $N_m = 22.5$). Let us consider the case of the location of the relay station on the edge of the lunar disk and the maximal distance of the Moon from Earth $R_{max} = 406,670$ km (apogee). In this case the aspect angle of Earth's surface from the Moon is equal to $\alpha = 1.8^{\circ}$. /200

We have noted that the time for a complete rotation of the Moon about its axis coincides with the time of revolution of the Moon about Earth, and therefore the Moon always turns the same side toward Earth. Thus, the transmitting antenna of the lunar relay station, once turned toward the Earth, will thereafter remain in that position, i.e., it is stabilized. The antenna stabilization accuracy is determined by the lunar libration and is about $\gamma = \pm 5^{\circ}$.

Thus, the width of the principal lobe of the transmitting antenna pattern of the lunar relay station in the case of natural stabilization is equal to $\theta = 12^{\circ}$.

In the case of artificial orientation of the antenna system along the radio beam of the ground station with the use of three stations equidistantly spaced along the equator of Earth, the stabilization accuracy is $\gamma = \pm 0.9^{\circ}$ and the width of the principal lobe of the transmitting antenna diagram of the lunar relay station can be taken equal to $\theta = 3.6^{\circ}$. Let us also consider the case of complete stabilization ($\gamma = 0^{\circ}$). Evidently the width of the main lobe in this case will be minimal and equal to $\theta = 1.8^{\circ}$.

Considering that the average directivity factor of an antenna pattern with width (to half-power points) of θ is equal to /201

$$G = \frac{2}{1 - \cos\frac{\theta}{2}},$$

and the geometric diameter of the antennas is equal to

$$D_1 = \frac{\lambda}{\pi} \sqrt{\frac{G}{\eta}},$$

where η is the antenna area utilization factor, equal approximately to 0.5, we perform the calculations for D_1 with $\lambda = \lambda_{opt}$ and $\theta = 12^\circ, 3.6^\circ, 1.8^\circ$.

The results of the calculations are summarized in tables II-3, II-4, II-5 (for $\theta = 12^\circ, 3.6^\circ, 1.8^\circ$, respectively).

Analysis of the data of the tables shows that with operation at /202 the optimal wavelengths the required dimensions of the receiving-transmitting antenna of the lunar relay station are quite large, particularly with the relay of telephony, where the optimal wavelengths are long.

TABLE II-2

Modulation Mode	Optimal wavelength, cm				
	$T_0 = 50^\circ K$	$T_0 = 100^\circ K$	$T_0 = 300^\circ K$	$T_0 = 600^\circ K$	$T_0 = 1{,}500^\circ K$
Television (F_0 = 94 Mcps)	15	15	15	15	15
Telephony (F_0 = 67 kcps)	40	45	55	57	60
PTM telegraphy (F_m = 1 kcps)	15	15	15	15	15

TABLE II-3

Modulation Mode	Antenna diameter, m				
	$T_0 = 50^\circ K$	$T_0 = 100^\circ K$	$T_0 = 300^\circ K$	$T_0 = 600^\circ K$	$T_0 = 1{,}500^\circ K$
Television	1.30	1.30	1.30	1.30	1.30
PTM telegraphy	1.30	1.30	1.30	1.30	1.30
Telephony	3.45	3.90	4.75	4.90	5.20

TABLE II-4

Modulation mode	Antenna diameter, m				
	$T_0 = 50^oK$	$T_0 = 100^oK$	$T_0 = 300^oK$	$T_0 = 600^oK$	$T_0 = 1,500^oK$
Television	4.35	4.35	4.35	4.35	4.35
PTM telegraphy	4.35	4.35	4.35	4.35	4.35
Telephony	11.60	13.00	16.00	16.60	17.40

TABLE II-5

Modulation mode	Antenna diameter, m				
	$T_0 = 50^oK$	$T_0 = 100^oK$	$T_0 = 300^oK$	$T_0 = 600^oK$	$T_0 = 1,500^oK$
Television	8.70	8.70	8.70	8.70	8.70
PTM telegraphy	8.70	8.70	8.70	8.70	8.70
Telephony	23.20	26.00	32.00	33.00	34.80

There are ways of reducing the required size of the lunar relay antenna. One of them is to shorten the operational wavelength with respect to the optimal value. Analysis of the plots of the relationship of $P_t/P_{t\ min}$ as a function of λ (fig. I-43) shows that the extremum of this function is not sharp, and the operational wavelengths can be chosen considerably shorter than the optimal values without a significant increase of the power required for the relay radiation. Thus, in the case of relay of telephony the changeover to an operational wavelength $\lambda = 15$ cm requires increase of the radiated power by no more than a factor of two. Hereafter, we shall assume that the operational wavelength for all modulation modes is $\lambda = 15$ cm. Consequently, the required antenna diameter for these three cases of lunar relay antenna stabilization will be, respectively, $D_1 = 1.3$ m, 4.4 m and 8.7 m.

Apparently the most promising antenna for the lunar relay station is the inflatable antenna made of synthetic material (metallized outer surface) of the mylar type (ref. 125).

Let us make an analysis of the required radiation power of the lunar radio relay station with operation at the $\lambda = 15$ cm wavelength with a ground antenna diameter $D_2 = 20$ m for the three cases.

1. Relay of an FM television signal with the following parameters

$N_0 = 16$; $F_0 = 94$ Mcps; $F_m = 5$ Mcps; $m = 8.4$; $N_m = 10^4$; $\nu_{inst} = 10^{-4}$;

2. Relay of an FM telephony signal having the following parameters

$N_0 = 16$; $F_0 = 67$ kcps; $F_m = 3$ kcps; $m = 11.1$; $\nu_{inst} = 10^{-4}$; $N_m = 10^4$;

TABLE II-6

θ	Radiation power P_t, w				
	$T_0 = 50^{\circ}K$	$T_0 = 100^{\circ}K$	$T_0 = 300^{\circ}K$	$T_0 = 600^{\circ}K$	$T_0 = 1{,}500^{\circ}K$
1.8°	$5.1 \cdot 10^{1}$	$5.9 \cdot 10^{1}$	$9.20 \cdot 10^{1}$	$1.42 \cdot 10^{2}$	$2.92 \cdot 10^{2}$
3.6°	$2.1 \cdot 10^{2}$	$2.4 \cdot 10^{2}$	$4.0 \cdot 10^{2}$	$9.8 \cdot 10^{2}$	$2.20 \cdot 10^{3}$
12°	$2.4 \cdot 10^{3}$	$2.8 \cdot 10^{3}$	$4.40 \cdot 10^{3}$	$6.80 \cdot 10^{3}$	$1.4 \cdot 10^{4}$

TABLE II-7

θ	Radiation power P_t, w				
	$T_0 = 50^{\circ}K$	$T_0 = 100^{\circ}K$	$T_0 = 300^{\circ}K$	$T_0 = 600^{\circ}K$	$T_0 = 1{,}500^{\circ}K$
1.8°	$1.88 \cdot 10^{-1}$	$2.2 \cdot 10^{-1}$	$3.4 \cdot 10^{-1}$	$5.0 \cdot 10^{-1}$	1.05
3.6°	$7.3 \cdot 10^{-1}$	$8.9 \cdot 10^{-1}$	1.3	2.0	4.2
12°	8.9	10.8	16.3	24.4	51.0

TABLE II-8

θ	Radiation power P_t, w				
	$T_0 = 50^{\circ}K$	$T_0 = 100^{\circ}K$	$T_0 = 300^{\circ}K$	$T_0 = 600^{\circ}K$	$T_0 = 1{,}500^{\circ}K$
1.8°	$1.5 \cdot 10^{-3}$	$1.8 \cdot 10^{-3}$	$2.8 \cdot 10^{-3}$	$4.2 \cdot 10^{-3}$	$0.90 \cdot 10^{-2}$
3.6°	$6.0 \cdot 10^{-3}$	$7.0 \cdot 10^{-3}$	$1.1 \cdot 10^{-2}$	$1.7 \cdot 10^{-2}$	$3.6 \cdot 10^{-2}$
12°	$7.0 \cdot 10^{-2}$	$8.5 \cdot 10^{-2}$	$1.4 \cdot 10^{-1}$	$2.1 \cdot 10^{-1}$	$4.3 \cdot 10^{-1}$

3. Relay of a telegraph signal transmitted using pulse-time modulation (PTM) and having the following parameters /203

$N_m = 22.5$; $F_m = 1$ kcps; $F_0 = 2$ Mcps; $M = 1000$ (duty cycle), $\nu_{inst} = 10^{-4}$.

The results of the calculations are summarized in tables II-6, II-7 and II-8, respectively.

From the previously presented consideration of the possibilities of the use of the various power sources as long-term operational onboard sources under the conditions of outer space, it follows that at the present only the solar cells satisfy the requirements. With the use of solar cells as the sources of electric energy for the supply of the radio equipment of the lunar relay station, the magnitude of the direct current power (and consequently the maximal power of the transmitter radiation) is determined by the value of the area of the solar cell panel. Assuming, as before, that the maximal value of the panel is /204
10 m^2, and the efficiency of the solar cell is 10 percent, we find that the overall efficiency of the relay station of 10 percent gives a maximal radiation power of the transmitter of about 100 W.

Assuming the presence of the relay station on the Moon, on the basis of the arbitrary numerical example presented above we can make the following conclusions.

1. Active relay of high-fidelity FM television without a tracking filter with use of lunar relay is possible only with utilization on Earth of high-sensitivity receiving devices (molecular and parametric amplifiers), large receiving antennas and a high degree of stabilization of the relay transceiver antenna.

For example, with $\theta = 3.6^{\circ}$ (orientation of the lunar antenna along the radio beam of one of the equidistant equatorial stations) and use of a parametric amplifier at the input of the ground receiver ($T_0 = 100^{\circ}K$) communication is possible only with the use of a ground antenna with $D_2 = 30$ m.

Use of a tracking filter at the ground receiver reduces the power requirement of the radio link by about a factor of 4. In our example the use of the tracking filter permits reduction of the ground antenna diameter to $D_2 = 15$ m (the antenna diameter of the lunar relay station $D_1 = 4.4$ m).

2. Active relaying of high-fidelity FM telephonic signals with the use of a radio relay station on the Moon, even without a tracking filter in the ground receiver, is possible with relatively small dimensions of the ground antenna.

For example, with $\theta = 3.6^{\circ}$, $T_0 = 300^{\circ}K$ communication is possible with the use of a ground antenna with diameter $D_2 = 2.2$ m.

With a tracking filter used in the ground receiver we can maintain communications even with a conventional crystal mixer at the receiver input with a ground antenna diameter $D_2 = 2.2$ m.

3. Active relaying of PTM telegraphy with the use of a relay station on the Moon is possible even with the use of simple receivers and small antennas on the ground.

Thus, in the case $\theta = 3.6^{\circ}$, $T_{0} = 1,500^{\circ}K$ communications are possible /205 with use on the ground of a simple antenna of the "wave duct" type (ref. 52).

The power margins available in the relaying of telephonic and telegraphic signals can be utilized to increase the number of channels (number of trunks in the relay station).

3. Passive Relaying of Signals Via the Moon

The first attempts to obtain reflected signals from the Moon were made before World War II in the U.S. by personnel of the Naval Research Laboratory, but they were unsuccessful. The first successful attempt at radiolocation of the Moon was made in Hungary in 1946, using military radar equipment. Later experiments were conducted on bouncing radar signals off the Moon in Australia, England and the U.S. The first Moon relay of a telegraphic message was accomplished in 1951, using a frequency of 418 Mcps between Cedar Rapids (Iowa) and Star King (Virginia) (ref. 84).

Relay of telephony signals via the Moon was accomplished in 1958 by personnel of the U.S. Naval Research Laboratory led by Trexler (ref. 85), and somewhat later by the team of the experimental station of Manchester University in Jodrell Bank, headed by Lovell.

The success of these experiments showed that there is a real possibility of using the Moon as a reflector for longrange communications.

Let us analyze the possibilities and the characteristics of a radio link with passive relay via the Moon. In order to formulate the requirements on the parameters of the radio link communication using lunar reflection, we must know the characteristics of the signal reflected from the Moon and the coefficient of reflection of the lunar surface.

Detailed study of the characteristics of the signal reflected from the Moon shows that it has several very characteristic features.

1. The existence of rapid fading with considerable amplitude and durations in the range of (0.1-2) sec. The amplitude distribution follows the Rayleigh law.

2. Existence of slow fading with durations in the range from 10 min /206 to 2 hours.

3. Existence of limitations in the passband.

4. Existence of signal delay by about 2.5 sec.

The fact that the slow and rapid fading are of different nature was established by Kerr and Shain in 1950 (ref. 86) as a result of their experiment on radar location of the Moon from Sydney. They also explained the mechanism of the rapid fading. They showed that because of the "roughness" of the lunar surface the reflected signal contains components of different relative phase shifts, and as a result of the libration (rocking) phenomenon the phase relations between these components change, the result being the rapid fadings which have received the name "libration fading."

It is of interest to know the amplitude of the fading resulting from the lunar libration. As we have mentioned, the distribution of the amplitudes of the reflected signal with account for only the rapid fading is Rayleighean. The distribution parameters depend on the magnitude of the libration and the frequency. Although no exact calculations on the distributions on the basis of the experimental data have been made, there is an indication (ref. 87) that in the course of 50 percent of the time of observation the amplitude fluctuations of the reflected signal amount to 4-5 db (power fluctuations of a factor of 2.5-3 times).

It is not known how the characteristics of the librational fading (in particular the amplitude distribution) vary with change of the operating frequency; relative to the rate of fading, it is known that it increases with increase of the frequency following an approximately linear law.

The maximal Doppler frequency shift due to the lunar libration is equal to

$$\Delta f_D = \pm\ f \iota_t \frac{R_M}{c}$$

where f is the carrier frequency;

ι_t is the lunar libration rate in rad/sec;

R_M is the radius of the lunar disk;

c is the speed of light.

Studies were made of the power spectrum P (f) of the signal reflected from the Moon, from which it was established that P (f) falls rapidly to 0 when

$$\frac{F_c}{F_{c_0}} > 0{,}3,$$ /207

where F_{c_0} is the spectrum width on the assumption that the Moon is a uniformly bright reflector (fig. II-1, ref. 88). From this it follows that the Moon has a very "dark" limb, that the effective scattering region is located in the center of the visible disk of the Moon and that the radius of this region amounts to about 1/3 of the lunar radius.

maximal rate of fading is approximately equal to

$$F_{fade} \approx \frac{2}{3} f \iota_t \frac{R_M}{c}$$

and for $\iota_t = 10^{-6}$ rad/sec and f = 3,750 Mcps (λ = 8 cm) is about 15 cps.

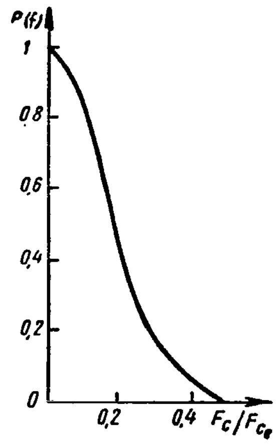


Figure II-1. Spectral distribution of power of signal reflected from Moon.

The cause of the slow fading was explained by Murray and Hargreaves of Joddrell Bank in 1954 (ref. 89). They showed that the slow fading arises with the reception of linearly polarized signals on a linearly polarized antenna as the result of the rotation of the plane of polarization of the wave with passage through the ionosphere (Faraday effect), and that it can be practically eliminated by the use of a receiving antenna with circular polarization.

The third characteristic of the signal reflected from the Moon is the frequency bandpass limitation. The cause of this phenomenon is the fact that the lunar surface is spherical rather than flat (fig. II-2).

If we assume that the Moon is a uniformly bright reflector clear out to its limb, a short pulse incident on the Moon should be stretched out to 11.6 msec, which corresponds to the time of travel of the radio waves of an additional 3,473 km (to the limb and back). Let us assume that we perform the relay of a telegraph signal using amplitude keying via the Moon. It /208
is evident that in this case the duration of the pulse (space) cannot be less than 11.6 sec, otherwise the pulses merge and no space will be noted. Thus, we find that communications are possible only with a modulation frequency of the order of 43 cps.

However, experimental studies have shown that the Moon is not a uniformly bright reflector.

Experiment shows that the main energy of the echo signal occurs in the first millisecond of its duration, and the majority of the energy is received in the first 100 msec (fig. II-3) (ref. 88). Thus, the experimental data obtained show that it is possible to build a communication system with passive lunar relay with a bandpass width to 10 kcps.

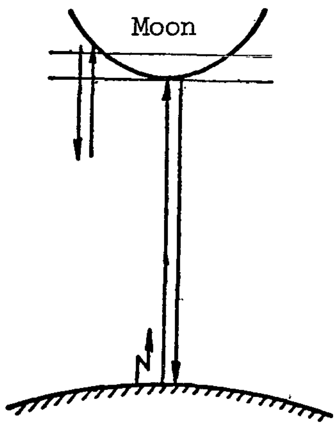


Figure II-2. Reflection of wave from lunar surface.

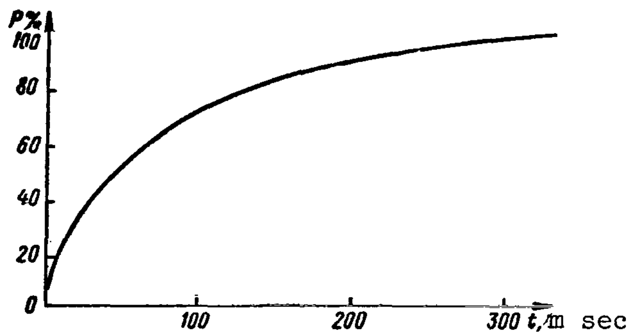


Figure II-3. Variation of relative distribution of energy P of signal reflected from Moon as function of time (in percent of total energy of pulse).

That portion of the reflected signal which contains the first 50 percent of the reflected energy corresponds to the first 8 km from the leading edge of the Moon, i.e., it is reflected by an area whose diameter is equal to 340 km, which is almost exactly one tenth of the lunar diameter.

In the reception of the signals obtained as the result of the /209 reflection of radiated pulses of short duration, the region of the rise of the reflected signal was always clearly outlined without any signs of the presence of small, previously arriving signals. This indicates that the region of the "sharp" reflection of the signals is the portion of the Moon which is closest to Earth and not some special formation somewhere on the lunar surface.

The Lambert law states that the reflected power is equal to

$$P_{\theta} = P_N \cos \theta,$$

where P_N is the power reflected along the normal to the surface;

P_θ is the power reflected at the angle θ to the normal.

In the present case, when the edges of the disk practically do not reflect, we can assume that the scattering law has the form

$$P_\theta = P_N(\cos\theta)^n, \text{ where } n = \text{const.}$$

In reference 88 numerical integration over the entire surface of the Moon was used to calculate the signal spectrum for a given value of n; this was then compared with the spectrum obtained as a result of observations. With n = 30 the difference between the spectra amounted to only a few percent. Thus, the lunar surface scattering law can be expressed by the equation

$$P_\theta = P_N(\cos\theta)^{30}.$$

This law assumes that the scattering takes place practically in the directions normal to the surface, and that only those elements of the surface which lie almost perpendicular to the incident ray (front edge of the lunar disk) reflect back to a significant degree. This corresponds to the assumption on the specular nature of the reflection, although it is difficult to imagine how the dielectric materials from which the lunar surface must be composed can be a specular reflector.

Reference 88 suggests that the theory of the existence of a lava surface, suggested by some selenologists, can be of assistance in the explanation of the observed form of the reflected signals in the experiments. In /210 this theory it is hypothesized that some time in the past lava flowed up to the surface of the Moon, then broke through and flowed along the valleys and craters. Here only the very highest peaks and ridges of the craters were above the surface of the lava sea, which then cooled and formed large areas over which only small irregularities appeared in the course of time. These large smooth surfaces can be the cause of the formation of the main reflected signal, on the background of which there are no random reflections from the mountains and the edges of the craters. In the power requirement calculations of the radio links with passive Moon relay it is necessary to know the reflective coefficient of the lunar surface. Different authors present somewhat differing values of this coefficient. In reference 88 it is indicated that the reflection coefficient of the lunar surface must be about $\rho_{eff} = 0.1$. According to the data of Evans (ref. 87), the coefficient of reflection is equal to only 0.03.

Let us turn to the calculation of the power requirement of the radio link using passive Moon relay.

For the selection of the optimal operating wavelength we analyze the communication equation (radar transmission equation)

$$P_t = \dot{N}_0 kT_{\text{eff}} F \frac{64 \cdot R^2 \cdot \lambda^2 \Gamma_1 \Gamma_2 N_1}{D_1^2 D_2^2 \pi S_0 \eta^2 \rho_{\text{eff}}},$$

where S_0 is the area of the lunar reflective surface;

ρ_{eff} is the reflection coefficient.

The notations of the other quantities are the same as before.

In the present case the margin factor N_1 must contain the margin coefficient for rapid fading of the signal reflected from the Moon, equal to 5 db, and also the margin coefficient for the polarization losses, equal to 3 db (ground reception is on a circularly polarized antenna and ground transmission is on a linearly polarized antenna). It is obvious that passive relaying of signals via the Moon is completely similar to passive relaying of signals using AES, and the optimal wavelengths for these two types of radio links will be identical. Thus, in the case of passive relay of signals /211 using the Moon, the optimal working wavelength will be $\lambda = 8$ cm ($f = 3{,}750$ Mcps).

Let us perform the power requirement analysis of a radio link with a passive Moon relay for the case of the transmission of discrete information with a rate of 1,000 baud ($F_m = 1$ kcps, high-speed telegraphy) and an operation using FM with frequency keying with $N_m = 10$. We take the diameters of the receiving D_{21} and transmitting D_{22} antennas as 20 m. We make the calculation for four values of the frequency instability coefficient $\nu_{inst} = 10^{-4}, 10^{-5}, 10^{-6}, 5 \cdot 10^{-7}$ and various T_0.

Let us consider that the maximal value of the frequency instability coefficient due to the Doppler effect is equal to $\nu_D = 6.2 \cdot 10^{-6}$. For the cases $\nu_{inst} = 10^{-6}$ and $5 \cdot 10^{-7}$ we assume that the Doppler frequency shift is compensated with an accuracy to 10 percent. We summarize the results of the computations in table II-9.

We see that the power consumption on the radio link using passive Moon relay with the transmission of high-speed telegraphy is quite considerable even with the use of high-sensitivity receiving equipment and large antennas at the receiving and transmitting ends of the radio link. For example, with the use at the ground station of a molecular amplifier with $T_0 = 50^\circ$K, $\nu_{inst} = 10^{-4}$ and $D_{21} = D_{22} = 20$ m, the radiative power required is 59 kW, which can

TABLE II-9.

$\nu_{inst} + \nu_D$	Radiation power P_t, kw				
	T_0=50° K	T_0=100° K	T_0=300° K	T_0=600° K	T_0=1500° K
$\nu_{inst} = 10^{-4}$, $\nu_D = 6.2 \cdot 10^{-6}$	59	69	108	166	340
$\nu_{inst} = 10^{-5}$, $\nu_D = 6.2 \cdot 10^{-6}$	30.5	35.6	56	84	174
$\nu_{inst} = 10^{-6}$, $\nu_D = 6.2 \cdot 10^{-7}$	18.4	21.5	34	108	227
$\nu_{inst} = 5 \cdot 10^{-7}$, $\nu_D = 6.2 \cdot 10^{-7}$	6.1	7.4	11.5	18.0	37.5

be provided only by parallel connection of four continuous-operation power klystrons.

With increase of the frequency stability, the power consumption is reduced. Thus, with $\nu_{inst} = 10^{-6}$ and with $\nu_D = 6.2 \cdot 10^{-7}$ and other conditions remaining the same, the radiative power required is 18.4 kW, which can be provided by the parallel connection of two continuous-duty klystrons. With still greater increase of the frequency stability, the condition /212

$$N_m \frac{2F_m}{F} > 1$$

is satisfied (strong signal) and linear detecting takes place (in this case filtering after the detector is equivalent to filtering ahead of the detector). The required radiative power P_t does not depend on F and is completely determined by the speed of operation F_m. In our case P_t = 6.1 kW, which is provided by a single continuous-duty klystron (ref. 79).

A considerable reduction of the power consumption of the radio link can be achieved by the use of diversity reception. Calculations show that with dual reception (reception on two separated antennas with automatic selection) the rapid-fading margin can be reduced by approximately a factor of two.

Reduction of the power requirements of the radio link can also be achieved by increase of the accuracy of pointing of the receiving and transmitting antennas at the Moon. The maximal possible gain which can be achieved is four. With high frequency stability, when the condition

$$N_m \frac{2F_m}{F} > 1$$

is satisfied (linear postdetection filtering), with reception on two separated antennas and a high accuracy of pointing at the Moon, the required radiated power is only 760 W. If the available power margin, equal to 20, is used to reduce the diameter of the antennas, antennas with $D_{21} = D_{22} = 10$ m are sufficient on the ground.

With high frequency stability, large receiving and transmitting antennas on the ground and with sensitive receiving equipment, even passive Moon relay of telephone signals is possible.

As a result of the frequency bandpass limitation due to the /213 sphericity of the reflector (the Moon), in the present case it is not possible to realize the advantages of the wideband modulation systems (FM and PCM). For this reason, for the transmission of analog information (telephony) in the present case it is most advisable to use the SSB system. With $D_{21} = D_{22} = 25$ m, $F_m = 3$ kcps, $N_m = 100$ (medium fidelity), single reception, $T_0 = 50^oK$ and the existence of a margin for inaccurate pointing at the Moon by the receiving and transmitting antennas (a factor of 4 with respect to power), the required transmitter radiation power $P_t = 40$ kW, which can be provided by parallel connection of three klystrons. With high accuracy of pointing of the antennas at the Moon and other conditions remaining the same, the radiation requirement is $P_t = 10$ kW, which can be provided by a single medium power klystron (ref. 79). Further reduction of the power consumption on the radio link with passive Moon relay of telephone signals can be obtained by means of conversion of the speech structure with subsequent reduction of the bandwidth (speech compression). According to reference 36, with full conversion of the speech structure, when only the characteristic features of the speech sounds are transmitted over the communications line, we can obtain a gain of a factor of 5 - 50 times in reduction of the frequency band occupied (a corresponding gain of a factor of 5 - 50 in power, if the SSB system is used).

Returning to the case of the transmission of discrete information, we should note that for the passive relay of signals via the Moon the frequency telegraphy system is not optimal. According to reference 90 the use in this case of a wideband system with noise-like signals (Rake type) can give a gain of the order of 20 db (a factor of 100 times in power) in comparison with frequency keying. The existence of such a significant advantage of the systems with noiselike signals forces us to turn concentrated attention to them for the development of communication lines with passive Moon relay.

This consideration permits us to draw the following conclusions.

1. The specific nature of the reflection of the radio signal from the lunar surface sets an upper limit to the frequency band occupied by the radio signal, the result being that it is only possible to accomplish passive relay of relatively narrow-band information with a spectrum width at /214

the radio frequency of less than 10 kcps; narrow-band telephony, high-speed telegraphy and facsimile.

2. The power consumption for passive relay of radio signals via the Moon is quite high and from the point of view of the possibilities of microwave-generating devices communication via the Moon is possible only with the use of large transmitting and receiving antennas (with diameters of the order of 20 m), high-sensitivity receiving equipment (parametric and molecular amplifiers), the use of high-stability frequency references ($\nu_{inst} \leq 10^{-6}$), compensation of the frequency shift resulting from the Doppler effect and the use of optimal modes of modulation.

3. For the compensation of the rapid fading of the signal reflected from the Moon it is advisable to use various forms of diversity reception (frequency or spatial diversity).

4. For the transmission of digital information by means of passive relay of the radio signals via the Moon the systems with noiselike signals (of the Rake type) must be considered promising.

4. Duration of Communication Sessions with Moon Relay

Knowledge of duration of the communication sessions with signal relay via the Moon is of interest in the case of both active and passive relay.

The time of simultaneous visibility of the Moon from two or more corresponding points on Earth can be determined on the basis of data on the position of the Moon published in the Astronomical Calendar for the corresponding year.

An idea of the value of this quantity can be obtained from the data on the maximal and minimal duration of the communication session via the Moon between London and various points on Earth's surface, presented in reference 87 and table II-10.

The data in this table refer to 1959. In the calculations of the time of simultaneous visibility from London and the corresponding points, use is made only of the time interval when the elevation angle of the Moon /215 above the horizon is greater than 7°.

Analysis of the data of table II-10 shows that the duration of the communication periods using Moon relay of the signals is in all cases considerable, adequate to transmit a large quantity of information.

For example, the maximal and minimal duration of the communication periods between London and Singapore (distance between points about 11,000 km) are, respectively, 6 h 55 m and 3 h 15 m. With telegraph operation at a rate of

TABLE II-10.

Station	Maximal duration of communication period per day		Minimal duration of communication period per day	
Calcutta	8 hours	30 min	3 hours	54 min
Adelaide	3	35	1	52
Central Nigeria	13	14	9	46
Montreal	10	18	4	05
Lusambo (Congo)	12	40	9	03
Capetown	11	57	9	59
Central Venezuela	9	46	5	53
Singapore	6	55	3	15

1000 baud (F_m = 1 kcps) during this time, we can transmit, respectively, one million words and 470,000 words. This rate of operation is provided even in radio links using passive relaying. In the case of active relaying, the information capacity of the communication channels using Moon relay can be increased by hundreds of times.

CHAPTER III. COMMUNICATION WITH SPACECRAFT WITHIN THE SOLAR SYSTEM

At the present time we are faced with the problem of providing /216 two-way radio communications between spacecraft directed from Earth to the planets of the solar system and Earth. We shall analyze the primary problems encountered in the resolution of this problem: the selection of the optimal operational wavelength, evaluation of the Doppler frequency shift, computation of the radio link power requirement, the capabilities of the various sources of supply for the onboard equipment, etc. But before turning to the analysis of these questions we shall briefly recall to the reader the basic data relating to the solar system itself.

1. Basic Data on the Planets of the Solar System

In the center of the solar system there is the Sun, about which there travel nine large planets, tens of thousands of small planets or asteroids, multitudes of comets and innumerable quantities of small meteoric particles. The total mass of all listed bodies is less than the mass of the Sun by a factor of about 750 fold. This is precisely what determines the central position of the Sun, having not only tremendous mass, but also gigantic dimensions. The diameter of the Sun is 1,390 thousand km. If we imagine Earth located at the center of the Sun, the Moon revolving about it would be far beneath the solar surface.

After the Sun, the most massive bodies of the solar system are the /217 planets.

The planets travel about the Sun in elliptic orbits. However, the eccentricities of the orbits of almost all planets of the solar system are quite small and do not exceed 0.1, except that for Mercury and Pluto they are somewhat larger than 0.2. Consequently, in the numerical evaluations of the planets the orbits can without exception be considered circular.

An interesting feature of the movement of the planets is the fact that they all revolve about the Sun in the same direction.

If we look at the system of planets "on edge," we can see that the planetary system is quite a flat formation: the orbits of six planets are inclined to the plane of Earth's orbit by less than 3.5°, and only for Mercury and Pluto are the inclinations somewhat larger (table III-1, ref. 91).

The basic data on the planets of the solar system and of the Sun are presented in table III-2, compiled on the basis of data of reference 54.

TABLE III-1

Planets	Inclination of planet orbital plane to plane of Earth's orbit
Pluto	$17^{\circ}8'$
Mercury	$7^{\circ}0'$
Venus	$3^{\circ}24'$
Earth	0°
Mars	$1^{\circ}51'$
Jupiter	$1^{\circ}18'$
Saturn	$2^{\circ}29'$
Uranus	$0^{\circ}46'$
Neptune	$1^{\circ}47'$

2. Selection of the Optimal Operational Wavelength

In the case of active transmission the communication equation is written in the form

$$P_t = N_0 k (T_0 + T_{ex}) F \frac{16R^2\Gamma_1\Gamma_2N_1}{G_1D_2^2\eta}, \qquad \text{(III-1)}$$

where P_t is the required radiation power of the onboard transmitter;

Γ_1, Γ_2 are the absorption coefficients of radio waves in the atmospheres of Earth and the corresponding planet; /219

N_0 is the s/n ratio at the detector input (RF);

T_0 is the equivalent noise temperature of the receiver itself;

T_{ex} is the equivalent external noise temperature;

F is the receiver bandpass width ahead of the detector (RF);

R is the distance between the communicating points;

G_1 is the spacecraft antenna gain;

D_2 is the ground antenna diameter;

η is the antenna area utilization coefficient ($\eta \approx 0.5$);

N_1 is the margin coefficient for polarization losses, losses in the antenna feed circuit, losses due to nonuniformity of the spacecraft antenna pattern, and losses due to the inaccuracy of the pointing of the ground antenna at the spacecraft.

TABLE III-2

Planets	Average planet diameter, km	Average orbital radius, millions of km	Time of rotation about axis	Time of revolution about sun	Maximal surface temperature, °K	Average velocity in orbit km/sec
Venus	12,200	108	No data	224.7 days	367	35
Mercury	5,000	57.9	87 days 23 hours	87.97 "	673	47.8
Mars	6,600	227.8	24 hours 37 min	686.98	303	24.1
Jupiter	139,500	777.8	9 hours 56 min	11.86 years	144	13.1
Saturn	120,000	1,426	10 hours 14 min	29.46 "	121	9.6
Uranus	47,000	2,869	10 hours 48 min	84.01 "	105	6.8
Neptune	43,000	4,495.6	15 hours 48 min	164.8 "	73	5.4
Pluto	5,800	5,929	No data	247.7 "	50	4.7
Earth	12,735	149.5	23 hours 56 min	365,266 days	333	29.8

Let us assume that $N_1 \approx 1$ (high antenna pointing accuracy, short feeder lines, matched polarizations of the onboard and ground antennas).

The conditions of the problem permit the introduction of several simplifying assumptions. It was established previously that the absorption of the radio waves in Earth's atmosphere is negligibly small, from 100 Mcps clear up to frequencies of 3,000 Mcps. Above 3,000 Mcps the absorption begins to increase as a result of the presence of resonance lines in the spectrum of the molecular components of Earth's atmosphere. However, the primary absorption occurs in the very lowest portion of Earth's atmosphere with a thickness of only a few kilometers. Thus, if the station for communication with interplanetary spacecraft is installed on the top of a mountain with a height of 5-6 km, the absorption of the radio signals in Earth's atmosphere will be small, up to high frequencies. Let us estimate, approximately, the lower limit of the operational range of wavelengths. Radiations of electromagnetic oscillations of high frequency (ultraviolet, X-rays) will be attenuated to a considerable degree with passage through Earth's atmosphere as a result of the losses due to ionization of the various components (molecular and /220 atomic) of the upper and lower atmosphere. It is known that the primary constituents of the upper atmosphere are nitrogen and oxygen in both atomic and molecular states (N_1, O_1, N_2, O_2 (refs. 92 and 61)). The values of the

ionization potential for the first level will be equal respectively, to 14.5; 13.5; 15.8; 12.5 ev (851; 781; 987; 735 Å). Thus, radiation with a wavelength of about 1,000 Å and shorter is attenuated to a significant degree during passage through the upper layers of the atmosphere. The ozone layer (its maximum lies at a height of 20-30 km, i.e., above the highest mountain peak on Earth) absorbs a significant portion of the ultraviolet radiation with a wavelength $\lambda \leq 3{,}000$ Å.

Consequently, by installing the communications station at a considerable height above sea level (5-6 km) we can extend the range of the operational wavelengths to the visible wavelength $\lambda \approx 4{,}000$ Å. We should note that in this case there is also a weakening of all refraction phenomena (reduction of the refraction angle and reduction of its fluctuations, since at the high frequencies the primary contribution is that of the tropospheric refraction, and particularly at low antenna elevation angles, when a significant portion of the path of the radio beam lies in the dense layers of the troposphere). The range of the operational wavelengths could be extended even further, if the communications station for the interplanetary spacecraft could be installed onboard artificial Earth satellites at a flight altitude of the order of 1,000 km. However, the power capabilities of the transceiving station on the AES will, considering the present state of the development of rocketry and electronics (limitation in weights, size and power), be considerably lower than at the ground station. We shall, therefore, consider the case of the installation of the transceiver station on an AES only as one of the possible future variants of the design of radio communication links with spacecraft.

We must keep in mind that other planets of the solar system have an atmosphere. For this reason our conclusion on the possibility of extension of the range of operational wavelengths up to $\lambda = 4{,}000$ Å is valid only when the spacecraft with the transceiving station is a satellite of the planet /221 or is on a sufficiently high point of the surface of the planet to neglect the effect of its atmosphere.

The second simplifying assumption is that we exclude from consideration those moments of time when the Sun or any other discrete source of radio noise falls in the ground-receiving antenna pattern. Thus, the only forms of external noise are the galactic noises and the noise of the planet with which communication is being established. We noted previously that the effective noise temperature in the antenna due to the galactic noise in the case of pointing of the receiving antenna toward the center of the Galaxy can be approximated by the following expression

$$T_{a\,\mathrm{gal}} \approx \lambda^{2,4} \cdot 469^\circ \mathrm{K}, \qquad \text{(III-2)}$$

where λ (wavelength) is measured in meters and $T_{a\ \mathrm{gal}}$ is measured in degrees Kelvin.

The effective noise temperature in the antenna due to the planet noises (planet surface temperature $T = T_{\mathrm{pl}}$) is equal to

$$T_{a\ \mathrm{pl}} = \left(\frac{\theta_1}{\theta_2}\right)^2 T_{\mathrm{pl}} \quad \text{for } \theta_1 < \theta_2, \qquad \text{(III-3)}$$

where θ_1 is the angular dimension of the planet with observation from Earth, equal to D_i/R_{av} (ratio of the planet diameter to the average distance between Earth and the planet), and θ_2 is the ground antenna half-power point pattern width, in the best case equal to $\theta_2 \approx \lambda/D_2$ rad. If $\theta_1 > \theta_2$, $T_{a\mathrm{pl}} = T_{\mathrm{pl}}$.

Considering all we have said, the communication equation can be written in the form

$$P_t = N_0 k \left[T_0 + \lambda^{2,4} \cdot 469 + \frac{\theta_1^2 D_2^2}{\lambda^2} T_{\mathrm{pl}}\right] F \frac{16 \cdot R^2}{G_1 D_2^2 \eta}. \qquad \text{(III-1')}$$

We should keep in mind that at the present stage of the development- /222 ment of the high-frequency radio wave generation technology the writing of the communication equation in the form (III-1') is, generally speaking, not entirely accurate. The reason is that the final objective of the analysis of the radio link is the determination of the required direct current power P_0, which is associated with the radiated power by the relation

$$P_t = P_0 \xi,$$

where ξ is the generator efficiency. At the present time ξ is strongly dependent on λ. In the range up to 1,000-3,000 Mcps ξ is approximately constant and amounts to only about 10-15 percent(ref. 79). With further increase of the frequency, ξ diminishes and in the millimeter wavelength range does not exceed 1 percent. Consequently, in the analysis of the radio link we write the communication equation relative to P_0 and include a factor showing the variation of ξ with wavelength.

However, as a result of several advantages in the millimeter wavelength region in comparison with the centimeter and decimeter ranges (higher accuracy in the coordinate determination with use of the millimeter wavelengths for radar, smaller antenna dimensions, etc.), at the present time scientists and engineers of various countries are making a major effort to increase the efficiency of the known methods of generation. As a result of this work we already note a tendency to relax the dependence of ξ on λ and to equalize the value of ξ in the ranges of the centimeter and millimeter wavelengths (ref. 79). Therefore, we shall assume ξ = const and shall consider the communication equation in the conventional form (in terms of P_t).

We derive the equations for the determination of λ_{opt} for several characteristic cases which may be encountered in practice.

Omnidirectional Antenna on Board the Spacecraft. Operation with an omnidirectional antenna on board a spacecraft located in the vicinity of any planet of the solar system is unfavorable from the power consumption viewpoint and will apparently be utilized only in emergency cases: if the ground- /223 tracking station loses the spacecraft, in the case of orientation of the directive antenna of the spacecraft with respect to the radio beam of the ground station; if the system for orientation of the spacecraft in space malfunctions (loses Earth), etc.

Here, in turn, we can consider two characteristic cases:

1. Strong signal

$$N_m \frac{2F_m}{F} > 1, \qquad (*)$$

where N_m is the receiver output s/n ratio;

F is the RF passband of the receiver;

F_m is the LF passband of the receiver.

With a limited power radiation of the onboard transmitter, this relation is realized in the case of almost complete compensation of the Doppler frequency shift and of the various instabilities, for example, by means of auto-tuning the heterodyne frequency of the ground receiver with respect to the arriving signal or by means of programmed variation of the heterodyne frequency.

As a result of the low effectiveness of the spacecraft transmitting antenna, it makes sense to consider only telegraphic operation with low rate. For definiteness let us assume that transmission is accomplished using frequency keying with a rate of 5 baud (F_m = 5 cps) and N_m = 10 (error probability about 10^{-4}).

Assuming that the RF passband (at the intermediate frequency) F is equal to triple the passband width of the channel filter (ref. 21), and that the passband width of the channel filter F_k is equal to

$$F_k = 2F_m + (\nu_{inst} + \nu_D) f_0 \tag{III-4}$$

(ν_{inst} and ν_D are coefficients characterizing the frequency shifts due to instabilities and Doppler effect), we find that the condition (*) is satisfied in the case when

$$(\nu_{inst} + \nu_D) f_0 < \frac{2F_m (N_m - 3)}{3}. \tag{III-5}$$

In the case F_m = 5 cps and N_m = 10 we find the additional broadening /224 of the passband of the channel filter due to the instabilities and the Doppler effect must not exceed 20 cps.

Knowing the operating frequency of the radio link f_0, we can find the required accuracy of compensation for the frequency shift ($\nu_{inst} + \nu_D$). With satisfaction of the condition (*) and use of a linear detector, the s/n ratio at the receiver input (at the limiter input) N_0 will be determined by the approximate equation

$$N_0 \approx N_m \frac{2F_m}{F}, \tag{III-6}$$

and the communication equation is written in the form (with $\theta_1 < \theta_2$)

$$P_t = 2N_m F_m \left[T_0 + \lambda^{2.4}\cdot 469 + \frac{\theta_1^2 D_2^2}{\lambda^2} T_{pl}\right]\cdot \frac{16\cdot R^2}{D_2^2 \eta}. \qquad \text{(III-7)}$$

Differentiating P_t with respect to λ, we obtain the equation for the determination of the optimal wavelength

$$4.4\,(\log \lambda_{opt}) = \log\left(\frac{\theta_1^2 D_2^2 T_{pl}}{565}\right) \qquad \text{(III-8)}$$

Relative to D_2 (ground antenna diameter) P_t is (with constant λ) a monotone decreasing function. Consequently, from the point of view of reducing the power consumption of the radio link, the diameter of the ground antenna must be increased. However, as a result of the limitation of the antenna fabrication accuracy the ground antenna diameter cannot be increased without limit, while retaining a constant operating wavelength. With some value $D_2 = D^*_2$ the operating wavelength λ becomes a critical wavelength for a given antenna, and with further increase of D_2 the performance of the antenna (efficiency, side lobe levels) decreases sharply. The critical wavelength for a given diameter D_2 is found from the condition

$$\lambda^*_2 = 16\cdot D_2\sigma, \qquad \text{(III-9)}$$

where σ is the relative accuracy of the fabrication of the antenna surface, equal to

$$\sigma = \frac{\lambda}{16\cdot D},$$

In the majority of the cases the analysis of the radio link is made /225 on the basis of a given diameter of the ground antenna D_2. In this case we make use of equation (III-8) for the determination of the optimal operational wavelength λ_{opt}. If, in this case, we find a value of $\lambda_{opt} < \lambda^*_2$ (critical wavelength for the given diameter of the ground antenna D_2), then we take

$\lambda = \lambda^*_2$ as the operational wavelength. If, however, the value obtained for $\lambda_{opt} > \lambda^*_2$, we take $\lambda = \lambda_{opt}$ as the operational wavelength found by solution of equation (III-8).

2. Weak signal

$$N_m \frac{2F_m}{F} \ll 1. \qquad (**)$$

This case is realized when the frequency shifts, due to the various instabilities, and the Doppler effect are not compensated or are only partially compensated.

In this case it is advisable to use a quadratic detector. It is known that then

$$N_0 \approx 1{,}27 \cdot \sqrt{\frac{4}{3} N_m \frac{F_m}{F}}. \qquad \text{(III-10)}$$

Consequently, the communication equation has the form

$$P_t = 1{,}27 \sqrt{\frac{4}{3} \cdot N_m F_m F}\, k \left[T_0 + \lambda^{2,4} \cdot 469 + \frac{\theta_1^2 D_2^2 T_{pl}}{\lambda^2}\right] \frac{16R^2}{D_2^2 \eta}, \qquad \text{(III-11)}$$

where $F \approx 3f_0\ (\nu_{inst} + \nu_D)$.

With a given diameter of the ground antenna D_2 the wavelength, which is optimal from the point of view of power requirements, is found from the condition of the minimum of expression (III-11). The equation for the determination of λ_{opt} has the form

$$\lambda_{opt}^{2,4} \cdot 1780 = T_0 + \frac{5\theta_1^2 D_2^2 T_{pl}}{\lambda^2} \qquad \text{(III-12)}$$

If the resulting value of $\lambda_{opt} < \lambda^*_2$ for a given value of the diameter of the ground antenna D_2, we take as the operational wavelength, as in the

preceding case, $\lambda = \lambda^*_2$. If, however, $\lambda_{opt} > \lambda^*_2$, then we take the /226
value of λ_{opt} obtained as the operational wavelength.

Highly Directive Spacecraft Onboard Antenna. The maximal value of the gain of the spacecraft antenna or ground transmitting antenna is limited only by the antenna diameter (ref. 52)

$$G_1 \leqslant G_{1_{max}} = \frac{\pi^2 D_1^2 \eta}{\lambda^2}. \qquad \text{(III-13)}$$

Substituting the value of $G_{1\ max}$ into the communication equation, we find

$$P_t = N_0 k \left[T_0 + \lambda^{2.4}\cdot 469 + \frac{\theta_1^2 D_2^2}{\lambda^2} T_{pl}\right] F \frac{16 R^2 \lambda^2}{\pi^2 D_1^2 D_2^2 \eta^2}. \qquad \text{(III-14)}$$

We see that the required radiated power P_t diminishes with reduction of the operational wavelength. However, we cannot conclude from this that the shortest wavelengths which can be used on a given radio link are optimal from the point of view of power requirements. This is because we have not taken into account one factor--the limitation in the accuracy of the fabrication of the antenna mentioned before. By reducing the operational wavelength λ with constant diameters of the ground and onboard antennas D_2 and D_1, we finally arrive at the situation where the wavelength becomes critical for one of the antennas: $\lambda^* = 16\sigma D_{1,2}$ (most probably for the ground antenna, since it can have larger dimensions than the onboard antenna). Further shortening of the wavelength can be achieved only under the condition that the diameter of the corresponding antenna is also reduced, i.e., under the condition that the ratio $\lambda/D = 16\sigma$ is held constant. Consequently, the communication equation for the wavelengths longer than the critical wavelength for the smaller of the antennas can be rewritten in the form

$$P_t = N_0 k \left[T_0 + \lambda^{2.4}\cdot 469 + \frac{\theta_1^2 T_{pl}}{256\cdot\sigma^2}\right] F \frac{(16)^3 R^2 \sigma^2}{\pi^2 \eta^2 D_1^2}. \qquad \text{(III-15)}$$

In this case the power potential (by power potential we mean the /227
product $P_t G_1$) of the radio link is maximal, and therefore we also include in our consideration the case of the transmission of television and telephone signals, using the method of frequency modulation, for which, as before, we

take $N_m = 10^4$ (high fidelity), $N_0 = 16$, $F_m = 5$ Mcps and 3 kcps; the spectrum widths of the radio signals F_0 will be 94 Mcps and 67 kcps, respectively.

As a result of the frequency shifts due to the various instabilities and the Doppler effect, the required RF receiver passband width F is equal to

$$F = F_0 + (\nu_{inst} + \nu_D)f_0.$$

Consideration of two characteristic cases is of interest

(a) $F_0 \quad (\nu_{inst} + \nu_D)f_0$ with $\lambda^*_1 < \lambda = \lambda^*_2$.

In this case the communication equation has the form

$$P_t = N_0 k\left[T_0 + \lambda^{2.4}469 + \frac{\theta_1^2 T_{pl}}{256\sigma^2}\right] F_0 \frac{(16)^3 R^2 \sigma^2}{\pi^2 \eta^2 D_1^2}. \qquad \text{(III-16)}$$

We see that the required radiation power of the onboard transmitter diminishes with a reduction of the operational wavelength. Consequently, the wavelength which is critical for the onboard antenna $\lambda_{opt} = \lambda^*_2 = \lambda^*_1$ will be optimal from the power requirement point of view. If antennas of identical manufacturing accuracy are used on the ground and on board, it is necessary to satisfy the condition $D_1 = D_2$ (or $\lambda_1^* = \lambda_2^*$). It is evident that we must try to use an onboard antenna of the largest possible diameter, since the radiative power required diminishes with increase of the diameter of this antenna.

However, the extremum of P_t is not sharp, and the required radiative power increases only slightly with λ larger than λ_{opt} (with retention of the ratio λ/D_2, i.e., with corresponding increase of the diameter of the ground antenna). Defining the width of the range of quasioptimal frequencies with with respect to a power increase of 10 percent in comparison with P_t for $\lambda = \lambda_{opt}$, we find that with $T_0 = 50^\circ$K (molecular amplifier at the input of the ground receiver) the lower limit will be $\lambda_{max} \geq 17$ cm, where the /228
equality sign holds with neglect of the term $\frac{\theta_1^2 T_{pl}}{256 \cdot \sigma^2}$ in comparison with $T_0 + \lambda^{2.4} \times 469^\circ$ K. In this case the dimensions of the ground antenna with

fabrication accuracies of $\sigma_1 = 10^{-4}$ and $\sigma_2 = 10^{-5}$ must be equal to $D_2 \geq 106$ m and $D_2 \geq 1060$ m, respectively. We see that the lower frequencies of the quasi-optimal range can hardly be realized in practice as a result of constructional difficulties in the creation of antennas of such gigantic dimensions.

(b) $F_0 \ll (\nu_{inst} + \nu_D)f_0$ with $\lambda^*_1 < \lambda = \lambda^*_2$.

In this case the communication equation is written in the form

$$P_t = N_0 k \left[T_0 + \lambda^{2,4} 469 + \frac{\theta_1^2 D_2^2}{\lambda^2} T_{pl}\right] (\nu_{inst} + \nu_D) f_0 \frac{16 R^2 \lambda^2}{\pi^2 D_1^2 D_2^2 \eta^2}. \quad \text{(III-17)}$$

We see that for a given D_2, in order to reduce the power consumption on the radio link, it is advisable to reduce the operational wavelength. The limit is the value $\lambda_2^* = 16\sigma D_2$.

Consequently, the wavelengths $\lambda_{opt} = \lambda^*_2$ will be optimal from the power consumption point of view under the condition that we use a ground antenna with the maximal possible diameter.

Setting the maximal diameter of the ground antenna equal to $D_2 = 100$ m, we find that the optimal wavelengths with antenna fabrication accuracies of $\sigma = 10^{-4}$ and 10^{-5} will be equal, respectively, to 16 and 1.6 cm. To obtain the maximal possible power gain, it is necessary to use on board the spacecraft a transmitting antenna with the maximum diameter D_1 possible for the onboard conditions.

The condition $F_0 \ll (\nu_{inst} + \nu_D)f_0$ can hardly be realized in practice in the transmission of telephonic and television signals using the method of frequency modulation with a high degree of modulation, even in the case of the absence of compensations for the frequency shift resulting from instabilities and the Doppler effect.

This condition (or its equivalent) is satisfied for the transmission of relatively narrow-band information by telegraphy. Therefore we shall consider still a third case (frequency keying). /229

(c) $$N_m \frac{2F_m}{F} \leqslant 1.$$

As we noted above, in this case the conditions

$$N_0 \approx 1.27\sqrt{\frac{4}{3} N_m \frac{F_m}{F}} \text{ and } F \approx 3(\nu_{inst} + \nu_D) f_0.$$

are satisfied. Then

$$P_t = \sqrt{4N_m F_m (\nu_{inst} + \nu_D) f_0}\left[T_0 + \lambda^{2.4} 469 + \frac{\theta_1^2 D_2^2 T_{pl}}{\lambda^2}\right]\frac{16R^2\lambda^2}{\pi^2 D_1^2 D_2^2 \eta^2}. \quad \text{(III-18)}$$

Here, just as in the case (b), the optimal wavelength from the point of view of power consumption will be $\lambda_{opt} = \lambda^*_2$, under the condition that a ground antenna with maximal possible diameter is used.

This analysis is not complete, since we have analyzed only the radio communication systems and have not considered the optical communication systems. With the appearance of the quantum-mechanical generators and amplifiers in the optical range (lasers), this range has become promising for use in space communications, and in certain cases for communication between spacecraft and ground stations (with location of the ground station high in the mountains or in the desert where cloud cover is slight).

The Laser on Board the Spacecraft. At the basis of the operation of the quantum mechanical devices--generators and amplifiers--lies the property of excited, coherently-radiating microparticles of matter to emit under definite conditions electromagnetic energy in the form of induced radiation of high power. This radiation excites oscillation in a cavity resonator tuned /230 to the frequency of the induced radiation.

As the cavity resonator in the quantum-mechanical devices use is made of a system of two parallel plates which form the so-called reflecting cavity resonator, in which the light waves undergo multiple reflections from the plates, which leads to the formation of standing waves. The distance between the plates (on the order of several centimeters) amounts to several thousand wavelengths of the optical and infrared bands. The reflecting plates must be ideally flat. The accuracy of the finish of their surface must amount to small fractions of wavelengths. This finish accuracy is achieved in the conventional optical devices (ref. 93).

Removal of the useful energy from the resonator is accomplished either through an opening in the plate (if it is not transparent) or through a plate made of semitransparent material. Thus, the radiation of the quantum-mechanical devices of the optical and infrared bands has very high directivity: a very narrow beam is formed in the device itself in the process of the generation of the oscillations. The light wave emerging from the instrument has initially a flat front. The angular width of the light beam is equal approximately to the ratio of the wavelength to the dimension (diameter) of the plates of the cavity resonator--about 0.1° (ref. 94). Further contraction of the beam can be accomplished with the aid of conventional optical lenses. In reference 95 it is indicated that the use of a 500 cm diameter telescope (maximal achievable diameter of the present telescopes) at a wavelength of $\lambda = 5000$ Å in conjunction with a laser makes it possible to obtain a beam width of 10^{-7} radians. A beam of this width will create on the Moon (average distance from Earth to Moon about 384,000 km) a circle of illumination with a diameter of only 40 m. A conventional projector would illuminate a circle on the Moon with a diameter of about 4000 km (the best projector forms a beam width of the order of 0.5°) (ref. 96). The accuracy of fabrication of the surface of the telescope, according to these data, is equal to

$$\sigma = \frac{\lambda}{16D} = 6.2 \cdot 10^{-9},$$

which is better than the existing fabrication precision of the surfaces /231
of the microwave antennas by more than four orders of magnitude.

We should note that the lasers have one essential deficiency--higher level of intrinsic noise than the analogous receiving devices of the radio frequency band (masers). This is because along with the induced emission there is spontaneous emission creating noise in the output signal in both generation and amplification regimes. Reference 97 presents comparative data on the sensitivity of the various receiving devices (table III-3).

Analysis of the data of table III-3 shows that the sensitivity of the optical receiving devices is considerably lower than the sensitivity of the receiving devices of the radio frequency band. Calculations show that /232
the equivalent noise temperature of the laser intrinsic noise is equal to $T_0 = 7.2 \cdot 10^4$ °K, while the minimal equivalent intrinsic noise temperature of the maser is equal to about 15°K. The spontaneous emission increases with increase of the frequency. Therefore the quantum mechanical devices cannot operate efficiently at the frequencies lying significantly above the ultraviolet range.

In the case of the use of the laser with variation of the operating wavelength λ in the limits $\lambda^*_1 < \lambda = \lambda^*_2$ (where λ^*_1, λ^*_2 — are the critical wavelengths for the smallest and the largest telescopes used), the communication equation is written in the form

$$P_t = N_0 k [T_0 + T_{pl}] F \frac{R^2 (16)^3 \sigma^2}{\pi^2 D_1^2 \eta^2}. \qquad \text{(III-19)}$$

Since with use on the ground of a telescope with mirror diameter of 5 m the width of the beam at the wavelength $\lambda = 5000Å$ is equal to $\theta_2 = 10^{-7}$ rad, even for communication with Pluto (maximal distance from Earth approximately $7.5 \cdot 10^9$ km), the linear width of the beam at the point where the spacecraft is located (on the planet) is less than the diameter of the planet. Consequently, as the temperature of the external noise T_{ex} we must take the temperature of the planet $T_{ex} = T_{pl}$.

Here we shall again consider two characteristic cases.

(a) $F_0 \gg (\nu_{inst} + \nu_d)f_0$ with $\lambda^*_1 < \lambda = \lambda^*_2$.

Let us try to picture the order of magnitude of $(\nu_{inst} + \nu_d)f_0$. Taking $\lambda = 5000Å(5 \cdot 10^{-5}$ cm), we find that $f = 0.6 \cdot 10^{15}$ cps = 600,000 Gcps. With $\nu_{inst} = 10^{-4}$ we have $\nu_\Sigma \cdot f_0 = 0.6 \cdot 10^{11}$ cps $= 60 \cdot 10^3$ Mcps. With the transmission of wide-band television, $F_0 \approx 10^2$ Mcps. Consequently, for the satisfaction of the condition $F_0 \gg \nu_{inst} f_0$ it is necessary that the frequency shift coefficient due to the instabilities and the Doppler effect be of the order of $\nu_\Sigma = 10^{-8}$ (this accuracy requires the compensation of the Doppler effect or the use of frequency heterodyne in the ground receiver in the case of autotuning of the frequency to the signal). With satisfaction of this condition, the required radiative power P_t does not depend on the wavelength in the range $\lambda^*_1 < \lambda = \lambda^*_2$ and the operational wavelength can be selected from considerations of convenience.

(b) $F_0 \ll (\nu_{inst} + \nu_d)f_0$.

This is the case of the transmission of wide-band and narrow-band /233
signals without compensation for the frequency shifts due to the Doppler effect and the instabilities.

In this case the required radiative power P_t is inversely proportional to the operational wavelength. Consequently, the longest wavelength of the range $\lambda^*_1 < \lambda = \lambda^*_2$, i.e., $\lambda_{opt} = \lambda^*_2$ is optimal from the power comsumption viewpoint. Thus, we must make the diameter of the ground telescope as large as possible.

TABLE III-3. OPTIMAL CHARACTERISTICS OF OPTICAL RECEIVING DEVICES

Receiving devices	Equivalent noise power/1 cps bandwidth, w	Optical frequency band, cps	Maximal modulation frequency, cps
Thermal Photoconductive Human eye	10^{-4} 10^{-13} 10^{-16}	10^{15} 10^{15} $3 \cdot 10^{14}$	-- -- 10
Photoemitter with subsequent maser amplifier	10^{-17}	10^{15}	10^{9} (max)
Photomultiplier With optical filter With optical heterodyne	10^{-17} 10^{-17} 10^{-17}	10^{15} 10^{12}(min) 10^{5} (min)	10^{8} 10^{8} 10^{8}
Optical maser pre-amplifier with heterodyne	10^{-18}	10^{5} (min)	10^{5} (min)
Incoherent amplifier photon sensor	10^{-18}	10^{9} (max)	10^{7} (max)
Microwave	$2 \cdot 10^{-22}$	10^{9}	10^{9}

Since the largest diameter of the current telescopes is 5 m, we shall assume that the optimal operational wavelength is $\lambda^*_2{}' = 5000$ Å. In this case the diameter of the telescope mirror on the spacecraft should also be made as large as possible to reduce the power consumption even further.

3. Analysis of Doppler Frequency Shift of Spacecraft Transmitter Located on Planetary Surface

For the determination of the Doppler frequency shift we must know the projection of the relative velocity v_1 of the communicating stations (in the present case Earth and the planet) on the radius vector drawn in the direction from one station to the other

$$\Delta f_D = v_1 \frac{f_0}{c}. \qquad \text{(III-20)}$$

It was shown previously that ν_1 is the derivative with respect to time t of the distance R(t) between the stations

$$v_1 = \frac{d}{dt} R(t).$$

It is evident that in this case the Doppler frequency shift is composed of two parts:

a slowly varying part, due to the revolution of the planet about the Sun;

a rapidly varying part due to the rotation of the planet about its own axis.

We shall determine these components separately. We shall make two assumptions which simplify the derivation of the equations considerably and reduce the accuracy only slightly:

(1) the planet orbits are circular; /234

(2) the planetary orbital planes coincide with Earth's orbital plane.

Let the angular velocities of travel of the planet and Earth about the Sun be equal to ω_1 and ω_2, respectively. Then, relative to the Sun-Earth radius vector, the angular velocity of the planet is equal to $\Omega = \omega_1 - \omega_2$, where $\Omega < 0$ for planets closer to the Sun than Earth and $\Omega > 0$ for planets more distant from the Sun than Earth, since all planets of the solar system revolve in the same direction (fig. III-1).

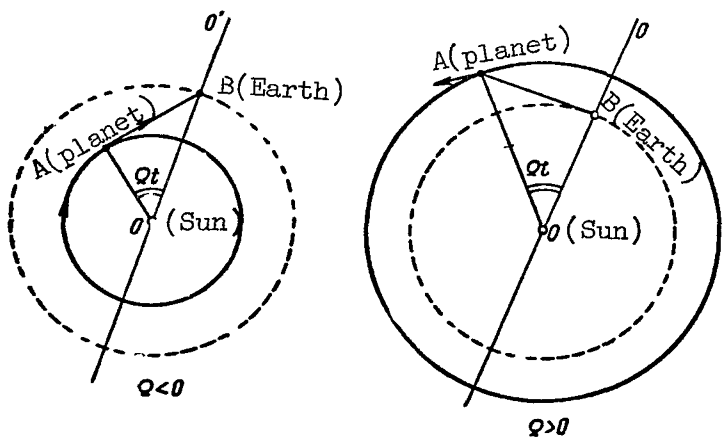


Figure III-1. Geometry of problem on determination of average distance between Earth and planet as function of time.

The slowly varying portion of the Doppler velocity is equal to

$$v_{11} = \frac{d}{dt} R_{av},$$

where R_{av} is the distance between the centers of Earth and the planet. Let us take as our reference the line OO' (case of closest approach of the two communicating stations).

From the triangle AOB using the cosine theorem we find /235

$$R_{av} = \sqrt{R_i^2 + R_e^2 - 2R_iR_e \cos \Omega t}, \qquad \text{(III-21)}$$

where R_i is the radius of the planet's orbit;

R_e is the radius of Earth's orbit.

Differentiating (III-21) with respect to time, we find that

$$v_{11} = \frac{R_iR_e \sin \Omega t\, \Omega}{\sqrt{R_i^2 + R_e^2 - 2R_iR_e \cos \Omega t}}. \qquad \text{(III-22)}$$

The rapidly varying portion of the Doppler velocity is equal to

$$v_{12} = \frac{d}{dt} R_t(t),$$

where R_t is the true distance between the points S_1 and S_2 on the surface of Earth and of the planet, with some average distance R_{av} between the centers of the planets.

For simplicity we shall consider that the axes of rotation of the planets are mutually parallel and are orthogonal to the plane of the eccliptic. In practice neither of these assumptions is satisfied. However, as a result of the fact that $v_{12} \ll v_{11}$, the error in the determination of the overall Doppler velocity $v_1 = v_{11} + v_{12}$ will be small.

The distance between the two points (S_1 and S_2) on the surface of Earth and the planet is equal to (fig. III-2)

$$R_t = \sqrt{(x_2 - x_1)^2 + (y_2 - y_1)^2}. \qquad \text{(III-23)}$$

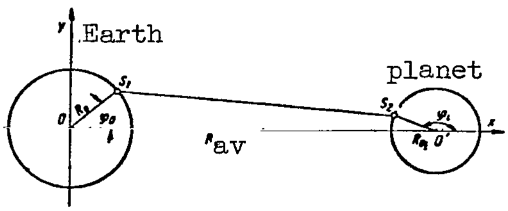


Figure III-2. Geometry of problem on determination of distance between points S_1 and S_2 on surfaces of Earth and planet as function of time.

From the figure we see that /236

$$\left.\begin{array}{ll} x_1 = R_0 \cos\varphi_0, & x_2 = R_{0_i} \cos\varphi_i + R_{av} \\ y_1 = R_0 \sin\varphi_0, & y_2 = R_{0_i} \sin\varphi_i, \end{array}\right\} \qquad \text{(III-24)}$$

where R_0 is the radius of Earth and R_{0_i} is the radius of the planet.

It is evident that $\varphi_0 = \omega_e t$, $\varphi_i = \omega_i t + \pi$, where ω_e and ω_i are the angular velocities of rotation of Earth and the planet, respectively.

The time t is reckoned from the moment of closest approach of the points S_1 and S_2. Differentiating (III-23) with respect to time, and using the fact that $R_{av} \gg R_{0_i}$ and $R_{av} \gg R_0$, we have, with a high degree of accuracy,

$$v_{12} \approx R_{0_i} \sin(\omega_i t + \pi)\,\omega_i + R_0 \sin(\omega_e t)\,\omega_e \qquad \text{(III-25)}$$

In practical calculations the accounting for the time dependence of both terms in the equation for the Doppler velocity v_1 is somewhat inconvenient because of the difference of the scales. Therefore, we shall consider the time dependence of only the first term (v_{11}) and assume the second term (v_{12}) to be constant and equal to the maximal value $v_{12\,max} = R_{0_i}\,\omega_i + R_0\,\omega_e$.

Thus, we perform the calculation of the Doppler frequency shift with variation of the relative position of the planets ($0 < \Omega t < 2\pi$) using the equation

$$\Delta f_D = 2\frac{f_0}{c}(v_{11} + R_{0_i}\,\omega_i + R_0\omega_e) = \gamma_D f_0. \qquad \text{(III-26)}$$

The maximal value of the Doppler frequency shift will be equal to

$$\Delta f_{D_{max}} = 2\frac{f_0}{c}(\Omega R_i + R_{0_i}\omega_i + R_0\omega_e). \qquad \text{(III-26')}$$

4. Calculation of the Basic Parameters of the Planet-Earth Radio Link

For the analysis of a specific radio link we must know the distance between the communicating points R_{av} (in the present case the average distance between Earth and the planet during a day-long period), the magnitude of the Doppler frequency shift and the angular dimension of the planet from Earth. The average daily distance between Earth and the planet is determined using equation (III-21). Let us compute $R_{av} = R_{av}(t)$ with variation of /238 Ωt in the range from 0 to 2π (synodic period of revolution of the planet about the Sun relative to an observer on Earth). The results of the calculation are shown in the curves of figure III-3.[1]

The frequency shift due to the Doppler effect is determined using /240 equation (III-26), in which it is considered that the frequency shifts may be of opposite sign.

Let us compute $\nu_d = \nu_d(t)$ with variation of Ωt in the range from 0 to 2π and present the results in the curves of figure III-4.

The exact values of the period of rotation about their own axes are not known for Venus and Pluto. For this reason the value of the Doppler frequency shift for these planets is not determined precisely. However, the error will be significant only in the determination of the minimal values of the Doppler frequency shift (with $\Omega t = 0, \pi$), and will be negligibly small in the determination of the Doppler frequency shift in any other cases of relative positioning of the communicating points. We determine the magnitude of the aspect angle of all planets of the solar system from Earth and present the results of the calculations in the curves of figure III-5.

We shall carry out the power requirement calculations for certain characteristic cases which may occur in practice.

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[1]All curves presented in the present chapter are identified with the corresponding planets using the following digits: 1, Venus; 2, Mars; 3, Mercury; 4, Jupiter; 5, Saturn; 6, Uranus; 7, Neptune; 8, Pluto.

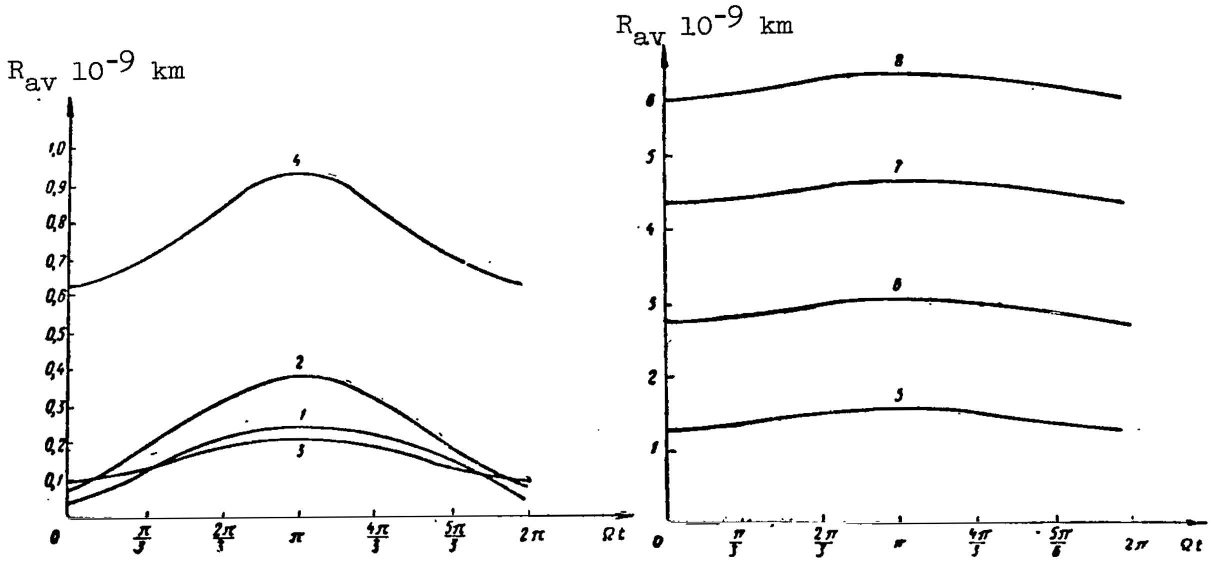

Figure III-3. Time variation of daily average distance from Earth to planet.

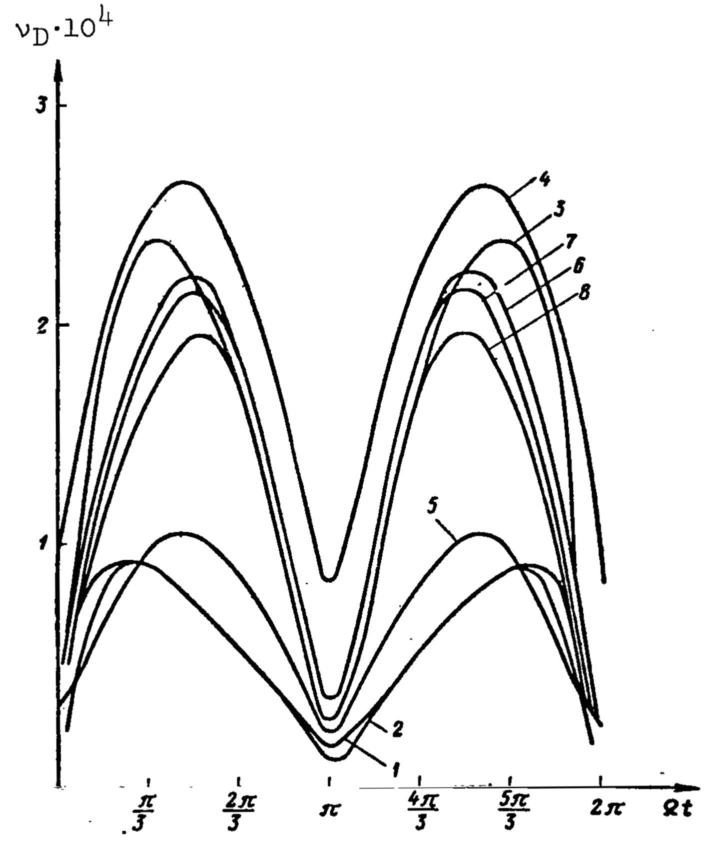

Figure III-4. Time variation of Doppler frequency shift (as fraction of carrier frequency f_0).

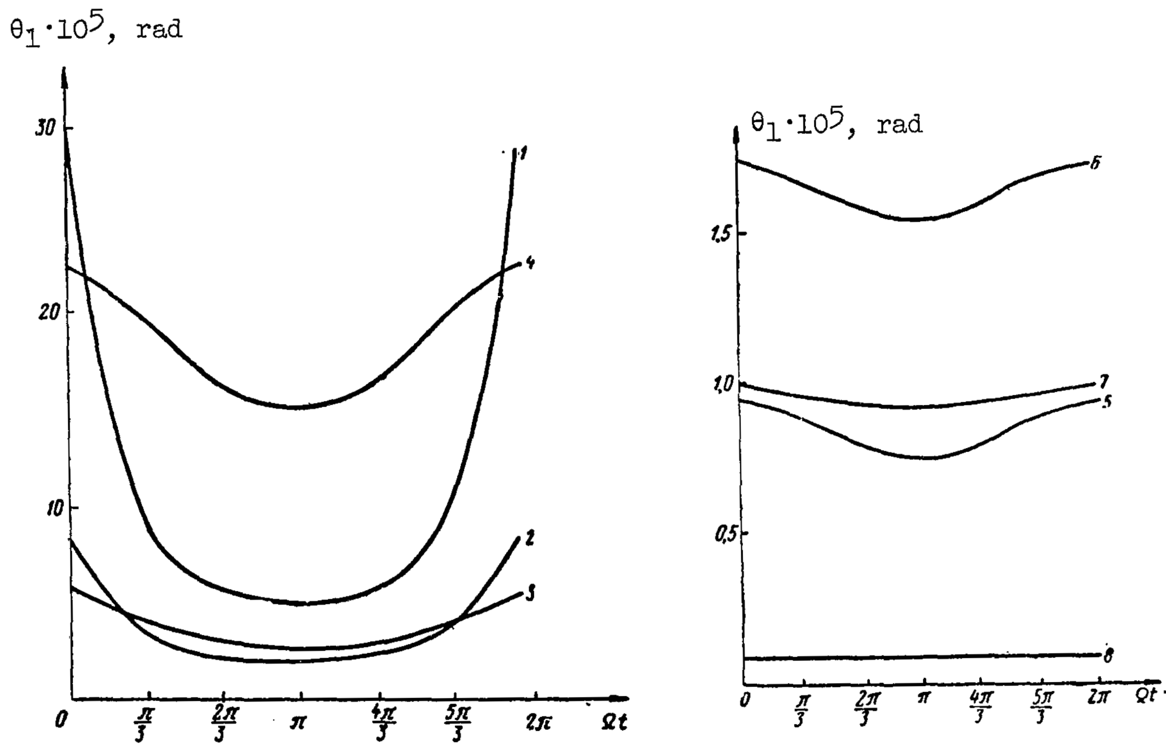

Figure III-5. Time variation of aspect angle of planet from Earth.

Omnidirectional Antenna on Board the Spacecraft.

(a) $N_m \frac{2F_m}{F} > 1$ (strong signal).

Assuming that $D_{21} = 25$ m and $D_{22} = 100$ m, and the antenna fabrication $\sigma = 10^{-4}$, we find λ_{opt_1} and λ_{opt_2} with variation of Ωt in the range from 0 to 2π. The results are shown in the curves of figure III-6a and b.

With an antenna fabrication accuracy $\sigma = 10^{-4}$ the critical wavelengths are: $\lambda^*_{21} = 4$ cm for $D_{21} = 25$ m and $\lambda^*_{22} = 16$ cm for $D_{22} = 100$ m. In those cases when $\lambda_{opt_1} < \lambda^*_{21}$ or $\lambda_{opt_2} < \lambda^*_{22}$, we take as the operating wavelength $\lambda = \lambda^*_{21}$ or $\lambda = \lambda^*_{22}$ (depending on the receiving antenna used).

We carried out the calculations of $P_t/P_{t\ min}$ for the determination of the width of the range of quasioptimal wavelengths (for the condition of increase of the required radiation power of the onboard transmitter by 10 percent) for the two extreme cases $\Omega t_1 = 0$ and $\Omega t_2 = \pi$ with $D_{21} = 25$ m and $D_{22} = 100$ m. /242

It was found that in all cases the range of variation of the optimal wavelengths $\lambda_{opt_1}(t)$ and $\lambda_{opt_2}(t)$ is within the range of the quasioptimal wavelengths, if as the lower boundary of the range of optimal wavelengths we take λ^*_{21} and λ^*_{22}. This indicates that there is no advantage in varying the operating wavelength as a function of the variation of the relative positioning of the communicating stations, and that it is only necessary to make a suitable choice of this frequency.

The wavelengths which are constant in the range $0 < \Omega t < 2\pi$, and which are best from the point of view of power consumption, and the values of the maximum power loss with operation at these wavelengths in comparison with operation at the optimal wavelengths $\lambda_{opt_1}(t)$ and $\lambda_{opt_2}(t)$ are presented in table III-4 ($D_{21} = 25$ m) and table III-5 ($D_{22} = 100$ m).

We carry out the calculation of the required radiation power of the onboard transmitter P_{t_1} and P_{t_2} at the operational wavelengths with the use of a ground receiver with a molecular amplifier at the input ($T_0 = 50^\circ$ K) and antennas with $D_{21} = 25$ m, $D_{22} = 100$ m, if transmission of a telegraphic /243

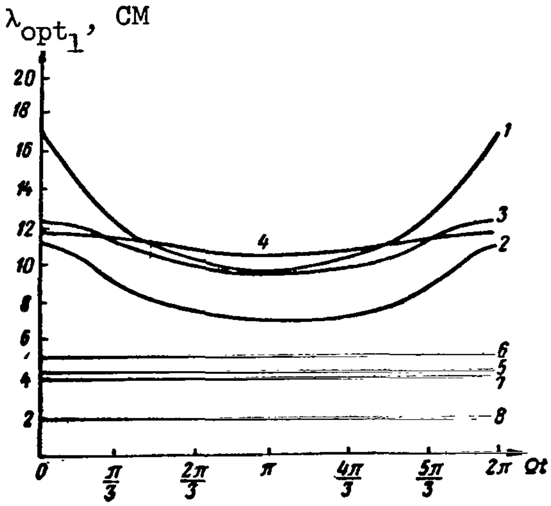

Figure III-6a. Variation of λ_{opt_1} with time (D_2 = 25 m).

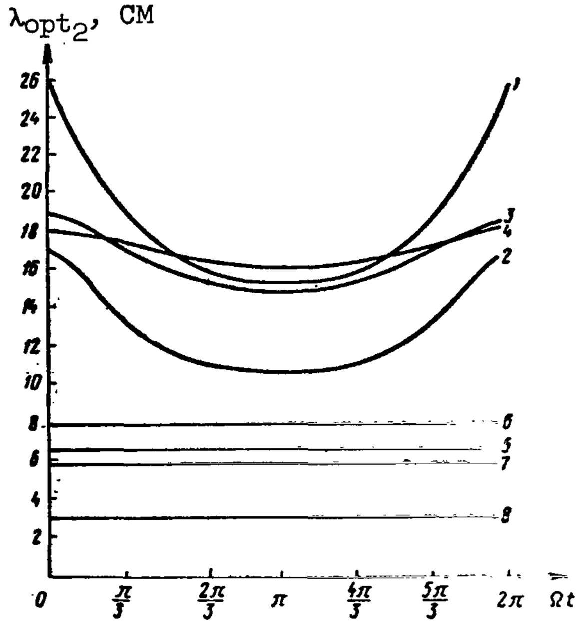

Figure III-6b. Variation of λ_{opt_2} with time (D_2 = 100 m).

TABLE III-4

Planet	Operating wavelength, cm	Maximal power loss in percent of $P_{t\ min}$
Venus	14	2
Mars	9.5	0.5
Mercury	11	0.4
Jupiter	11.5	0.2
Saturn	4.5	0.2
Uranus	5.2	0.2
Neptune	4	0.2
Pluto	4	0.2

signal (5 baud operating speed) is accomplished using frequency keying with a s/n ratio at the output N_m = 10 (low-frequency filter bandpass width F_m = 5 cps). The results of the calculation are shown in the curves of figures III-7 and III-8.

It was shown above that the condition

$$N_m \frac{2F_m}{F} > 1$$

TABLE III-5

Planet	Operating wavelength, cm	Maximal power loss in percent of $P_{t\ min}$
Venus	20	3.5
Mars	16	4
Mercury	16.7	0.4
Jupiter	16.7	0.2
Saturn	16	8.5
Uranus	16	6
Neptune	16	9
Pluto	16	14

in the case when $N_m = 10$ and $F_m = 5$ cps is satisfied with $\nu_\Sigma f_0 = (\nu_{inst} + \nu_d) f_0 < 20$ cps.

We calculate the values of ν_{Σ_1} and ν_{Σ_2} for the cases $D_{21} = 25$ m and $D_{22} = 100$ m and summarize the results in table III-6.

TABLE III-6

Planet	Required heterodyne frequency stability	
	$D_{21} = 25$ m	$D_{22} = 100$ m
Venus	$9.4 \cdot 10^{-9}$	$13.4 \cdot 10^{-9}$
Mars	$6.4 \cdot 10^{-9}$	$10.6 \cdot 10^{-9}$
Mercury	$7.30 \cdot 10^{-9}$	$11.2 \cdot 10^{-9}$
Jupiter	$7.66 \cdot 10^{-9}$	$11.2 \cdot 10^{-9}$
Saturn	$2.30 \cdot 10^{-9}$	$11.2 \cdot 10^{-9}$
Uranus	$3.5 \cdot 10^{-9}$	$10.8 \cdot 10^{-9}$
Neptune	$2.70 \cdot 10^{-9}$	$10.8 \cdot 10^{-9}$
Pluto	$2.70 \cdot 10^{-9}$	$11.0 \cdot 10^{-9}$

We see that in both cases a high stability of the frequency of the ground heterodyne is required, which is achieved only with the use of a molecular oscillator as a frequency standard. However, there is a way to reduce the requirements on the frequency stability of the ground heterodyne, which amounts to switching the operation of the radio link over the pulse operation (for example, pulse-time keying) with a high duty factor M (refs. 35, 21 and 43).

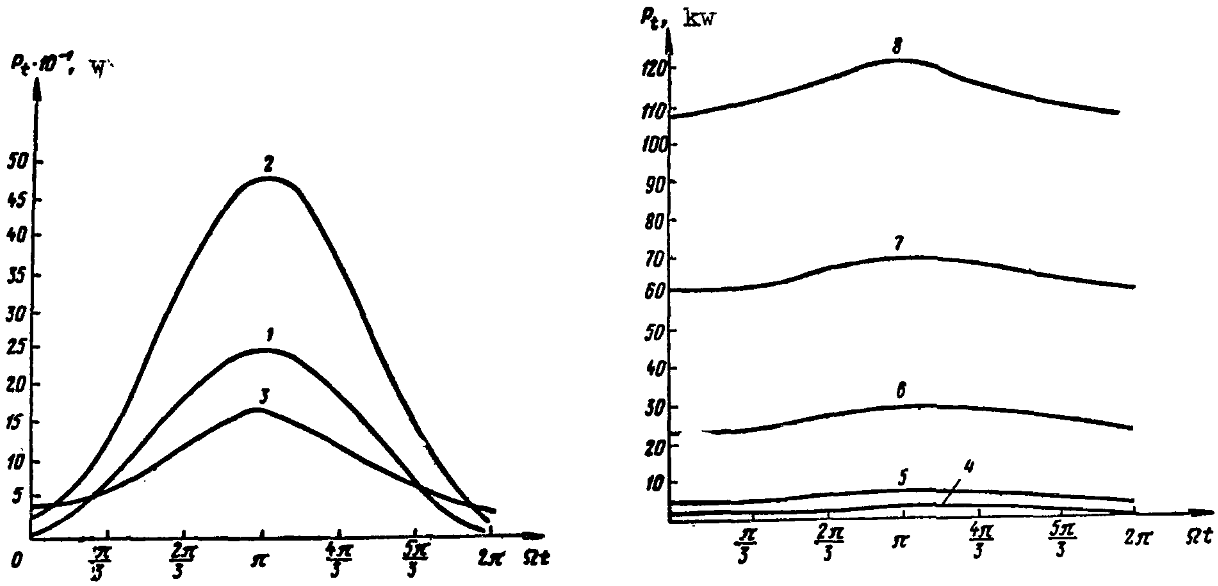

Figure III-7. Variation of radiated power of onboard trans-mitter with time for FT telegraphy, nondirectional antenna on board, $N_m \frac{2F_m}{F} > 1$ and $D_2 = 25$ m.

Since the width of the signal spectrum at the low frequency increases by approximately a factor of M times, both ν_{Σ_1} and ν_{Σ_2} increase by approximately M fold. With M > 1,000 the required stability of the frequency of the ground heterodyne can be provided by a quartz oscillator. /246

Analysis of the curves of figures III-7 and III-8 shows that even in the case of compensation of the frequency shifts, resulting from the instabilities and the Doppler effect, the power consumptions are significant. In the absence of compensation for the frequency shifts, the power consumptions will be higher by about n fold, where

$$n = \sqrt{\frac{2}{3} \frac{F}{2F_0 N_m}}.$$

In the case when f_0 = 1,000 Mcps, F_m = 5 cps, N_m = 10, $(\nu_{inst} + \nu_D) \approx 10^{-4}$, we find n = 50, i.e., the power requirements go up by almost two orders. Therefore, without carrying out detailed computations of the required radiation power P_t, we can say immediately that at the present stage of development it is not possible to maintain long-term radio communication with a

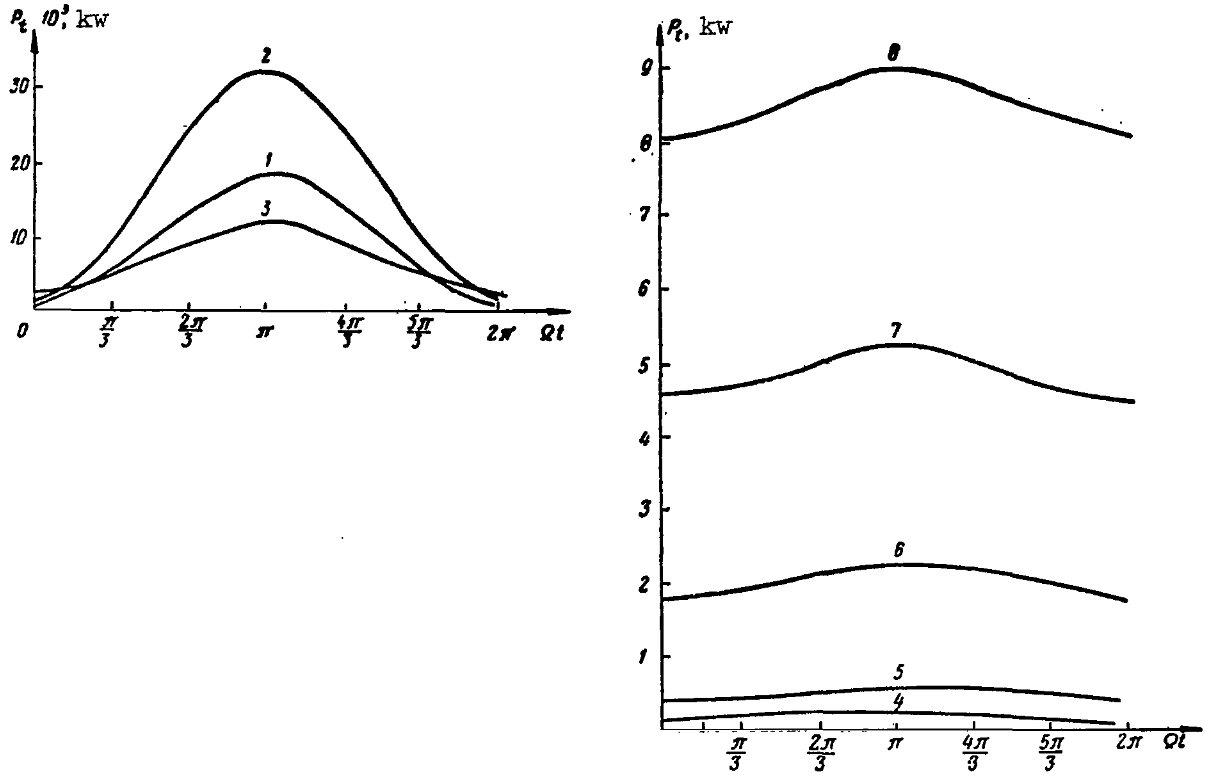

Figure III-8. Variation of radiated power of onboard transmitter with time for FT telegraphy transmission, nondirection onboard antenna, $N_m \frac{2F_m}{F} > 1$ and D_2 = 100 m.

spacecraft located in the vicinity of any planet of the solar system, if the spacecraft utilizes an omnidirectional transmitting antenna and steps are not taken to compensate the frequency shift due to the various instabilities and the Doppler effect (ref. 79).

Directional Antenna on Board the Spacecraft.

(a) $F_0 \gg \nu_\Sigma f_0$ with $\lambda^*_2 = \lambda > \lambda^*_1$.

The required radiation power of the onboard transmitter P_t depends to a considerable degree on the accuracy of the fabrication of the antenna (P_t is directly proportional to the square of σ). Therefore, one of the ways to reduce the power consumption on the radio link is the improvement of the accuracy of the fabrication of the ground and onboard antennas. As an example let us consider two values: $\sigma_1 = 10^{-4}$, $\sigma_2 = 10^{-5}$. With a diameter of the ground and onboard antennas D_2 = 5 m, the optimal wavelengths are, respectively, equal to λ_{opt_1} = 0.8 cm: λ_{opt_2} = 0.8 mm.

We make a calculation of the required radiation power of the onboard transmitter P_t with $\sigma_1 = 10^{-4}$ in the case of the transmission of a wideband television signal (F_m = 5 Mcps, N_0 = 16, $N_m = 10^4$) under the condition that a molecular amplifier with effective self-noise temperature $T_0 = 50^\circ$ K /247 is used at the ground receiver input. The results of the computations are shown in the curves of figure III-9. We also determine $P_{t_{max}}$ ($\Omega t = \pi$) for $\sigma_2 = 10^{-5}$.

For the case of the piloted spacecraft it is of interest to know the power consumption of the radio link with the transmission of speech. We shall make an analysis of the required onboard transmitter radiation power for the case of transmission of a telephonic signal using FM with F_m = 3 kcps, N_0 = 16, $N_m = 10^4$. The other parameters of the radio link are, as before, ($D_2 = D_1$ = 5 m, $\sigma_1 = 10^{-4}$, $T_0 = 50^\circ$ K). The results of the computation are shown in the curves of figure III-10.

Let us also determine $P_{t\ max}$ ($\Omega t = \pi$) with $\sigma_2 = 10^{-5}$.

The existence of a considerable delay of the response of the correspondent as a result of the finite velocity of propagation of the radio waves presents certain inconveniences for two-way telephonic conversation. Let us determine the range of variation of the response delay for various cases of location of the spacecraft. The results are summarized in table III-7.

TABLE III-7

Planet	Delay of correspondent's reply
Venus	4.6 min-28.8 min
Mars	8.7 min-41.6 min
Mercury	12.3 min-24.2 min
Jupiter	1 h 9.3 min-1 h 42.5 min
Saturn	2 h 22 min-2 h 55 min
Uranus	5 h 02 min-5 h 36 min
Neptune	8 h 03 min-8 h 36 min
Pluto	10 h 40 min-11 h 21 min

In order to satisfy the condition $F_0 >> \nu_\Sigma f_0$ (for example, by a factor of 10) with transmission of a television signal from the spacecraft, it is necessary that the frequency stability of the ground receiver heterodyne be equal to $\nu_{\Sigma_1} = 4.15 \cdot 10^{-4}$, $\nu_{\Sigma_2} = 4.15 \cdot 10^{-5}$ respectively, for σ_1 and σ_2.

For transmission of a telephonic signal the heterodyne frequency /250 stability must be higher: $\nu_{\Sigma_1} = 2.5 \cdot 10^{-7}$, $\nu_{\Sigma_2} = 2.5 \cdot 10^{-8}$ and, consequently, here a molecular oscillator apparently must be used as a frequency standard.

In the case of the transmission of telephonic (coded speech) and telegraphic signals it is advisable to switch to pulse operation of the channel with high duty factor M in order to reduce the frequency stability requirements.

In the present case the condition $F_0 \ll \nu_\Sigma f_0$ with $0 < \Omega t < 2\pi$ is not satisfied, even with transmission of the telephonic signal. (In the case of transmission of a wide-band television signal without compensation for the frequency shift resulting from instabilities and the Doppler effect, the opposite condition is satisfied: $F_0 \gg \nu_\Sigma f_0$.) Therefore, we shall not consider the transmission of continuous information (television, telephony) without compensation for the frequency shift, but shall limit ourselves in this case to the consideration of the transmission of discrete information.

(b) $N_m \frac{2F_m}{F} \ll 1$ (weak signal).

With $D_2 = 100$ m and $\sigma_1 = 10^{-4}$, $\sigma_2 = 10^{-5}$ we find $\lambda_{opt_1} = 16$ cm, $\lambda_{opt_2} = 1.6$ cm.

Let us make the computation (similar calculations were made in reference 121) of the required onboard transmitter radiation power P_t for $\sigma_1 = 10^{-4}$ at wavelength $\lambda = 16$ cm with $D_2 = 100$ m, $D_1 = 5$ m, $T_0 = 50^\circ$ K, if transmission is accomplished using frequency keying with a rate of 5 baud ($F_m = 5$ cps) and the frequency detector is built using the filter discriminator scheme, where the RF bandwidth is equal to three times the bandwidth of the channel filter F_k.

Let us consider two cases:

(1) the bandpass width of the channel filter of the ground receiver

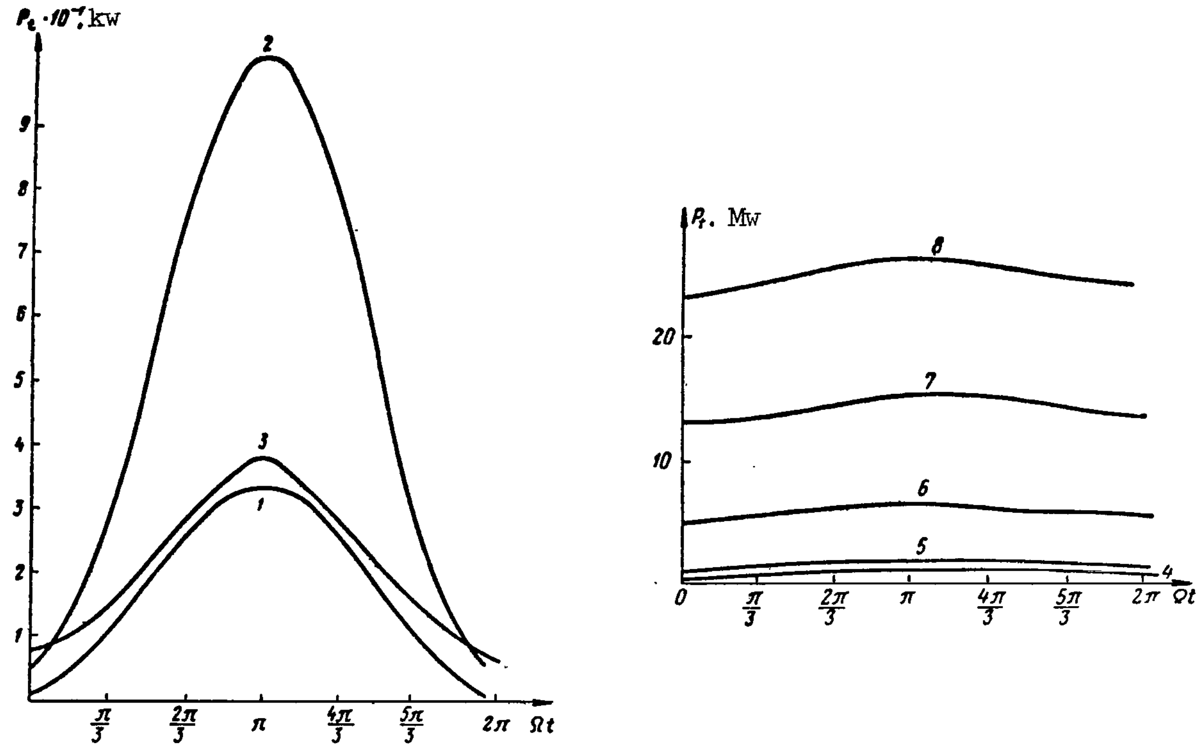

Figure III-9. Variation of radiated power of onboard transmitter with time for FM television transmission, $F_0 \gg \nu_\Sigma f_0$, $\sigma_1 = 10^{-4}$; $G_1 = G_{1\,max}$.

is constant and equal to the maximal frequency shift due to the Doppler effect and the various instabilities;

(2) the bandpass width of the ground receiver channel is variable and equal to the instantaneous frequency shift.

We calculate P_t for both cases ($F_k = F_{k_0}$ and $F_k = F_k(t)$) with /251
$\sigma_1 = 10^{-4}$ and present the results in the curves of figures III-11 and III-12.

Let us determine $P_{t\ max}$ for $F_k = F_{k_0}$, $\sigma_2 = 10^{-5}$ (with $\Omega t = \pi$).

The power consumption on the radio link can be reduced considerably with transfer of the channel to pulse operation with high duty factor M.

Let us determine gain C in the power requirement of the radio link with transfer of the channel from operation using frequency keying with $F_m = 5$ cps and $N_m = 10$ to operation using pulse-time keying with M = 10,000, $N_m = 10$

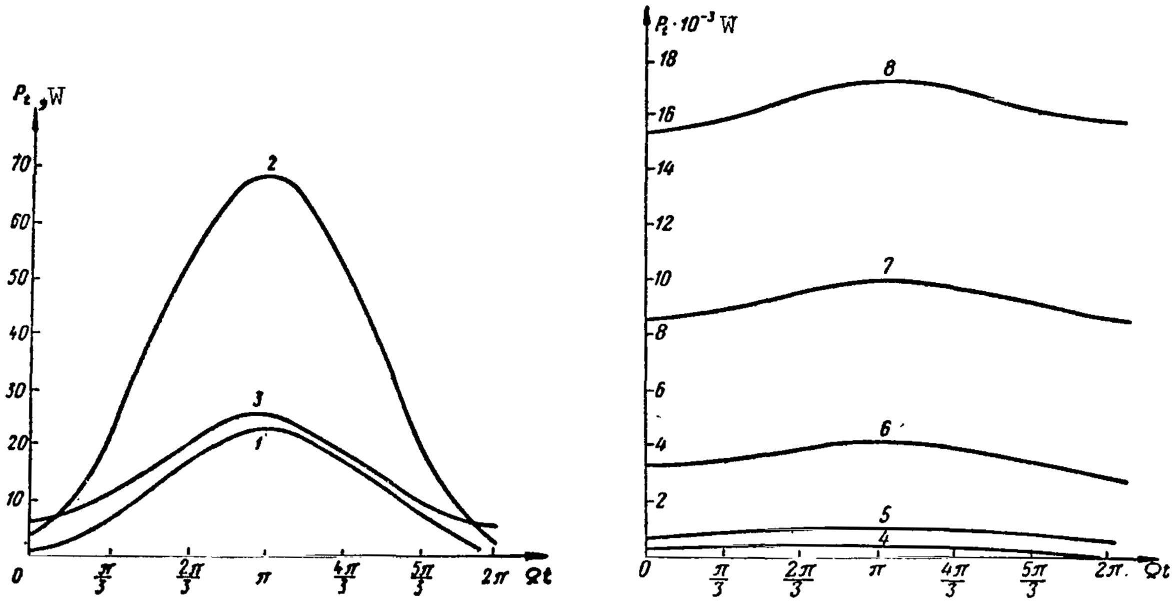

Figure III-10. Variation of radiated power of onboard transmitter for FM telephony transmission, $F_0 \gg \nu_\Sigma f_0$, $\sigma_1 = 10^{-4}$, $\bar{G}_1 = G_{1\,max}$.

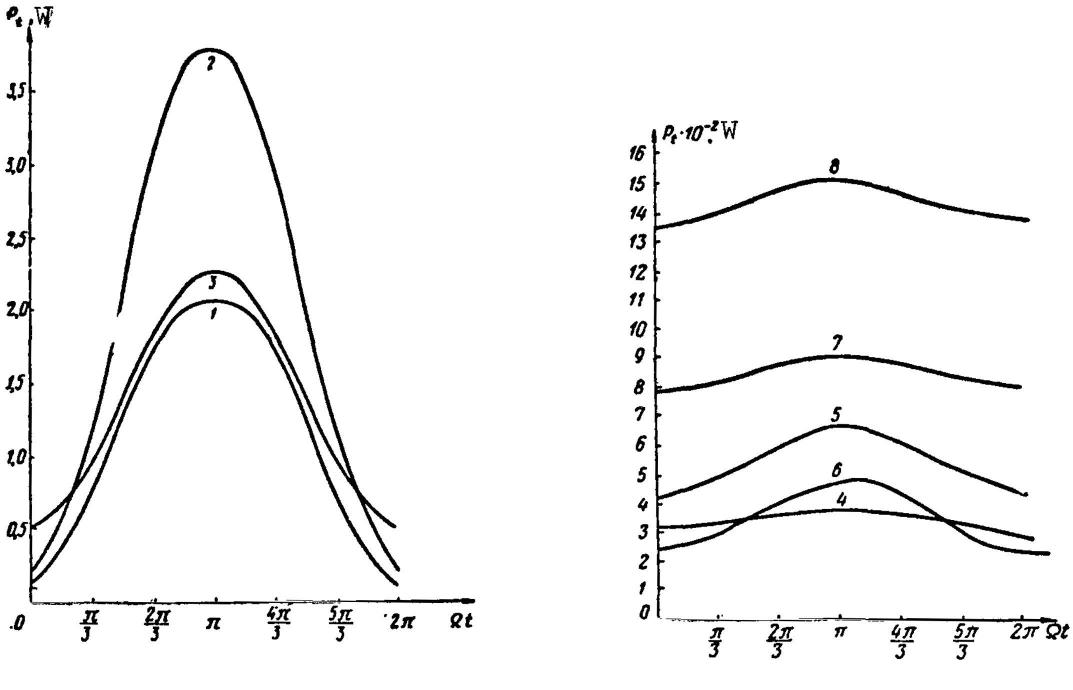

Figure III-11. Variation of radiated power of onboard transmitter for FT telegraphy transmission, $N_m \frac{2F_m}{F} \ll 1$, $F_k = \text{const}$.

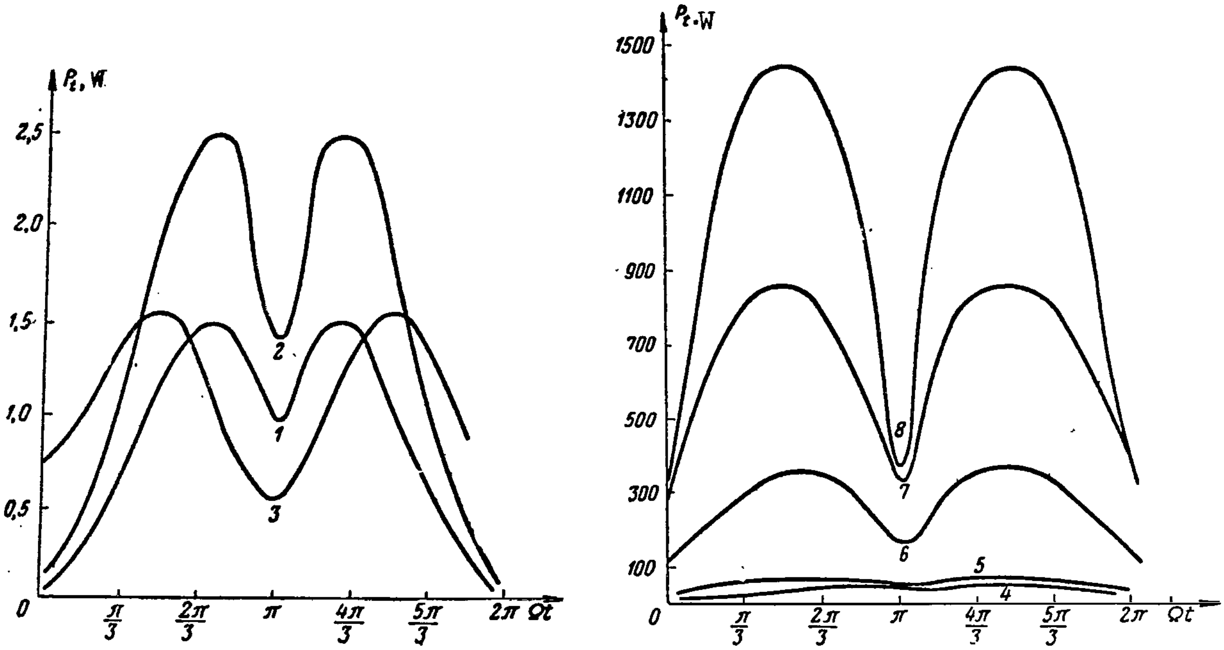

Figure III-12. Variation of radiated power of onboard transmitter for FT telegraphy transmission, $N_m \frac{2F_m}{F} \ll 1$, $F_k = F_k(t)$.

(same speed of operation), if synchronous reception of the PTM is used, the receiver being opened at the time of possible arrival of the pulses.

Let us compute C for the values $\sigma_1 = 10^{-4}$ and $\sigma_2 = 10^{-5}$ and the maximal Doppler frequency shift. We consider that in the case of frequency telegraphy $F_1 \approx 3\,(\nu_{inst} + \nu_d)f_0$, and in the case of PTM $F_2 = 2MF_m + (\nu_{inst} + \nu_d)f_0$.

We find that with $\sigma_1 = 10^{-4}$ the condition

$$N_m \frac{2F_m M}{F_2} > 1$$

is satisfied and in the case $\sigma_2 = 10^{-5}$ the condition

$$N_m \frac{2F_m M}{F_2} < 1.$$

is satisfied.

Consequently, in the case $\sigma_1 = 10^{-4}$ with reception of PTM a linear detector should be used; then

$$N_{0_2} \approx N_m \frac{2MF_m}{F_2},$$

and the gain will be equal to

$$C_1 \approx \sqrt{\frac{F_{k0}}{F_m N_m}}.$$

With $\sigma_2 = 10^{-5}$ in both cases (FT and PTM) a quadratic detector should be used, and we find $C_2 \approx \sqrt{M} = 100$ (we consider that $F_1 \approx 3F_2$).

We compute C_1 for various radio links and summarize the results of the calculation in table III-8.

TABLE III-8 /254

Planet	Power gain C	
	$\sigma_1 = 10^{-4}$	$\sigma_2 = 10^{-5}$
Venus	66.5	100
Mars	66.5	100
Mercury	103	100
Jupiter	107	100
Saturn	70	100
Uranus	98.5	100
Neptune	97.5	100
Pluto	94	100

We see that the use of the pulse modulation modes gives a considerable power gain.

Laser on Board the Spacecraft

(a) $F_0 \gg \nu_\Sigma f_0$ with $\lambda^*_2 = \lambda = \lambda^*_1$.

Let us carry out the calculation of the required radiation power of the onboard transmitter P_t in the case of the transmission of a high-fidelity television signal (F_0 = 94 Mcps, N_0 = 16) using FM (m = 8.4) if the diameters of the ground telescope and the onboard telescope are D_2 = 5 m and D_1 = 1 m, respectively, and operation is conducted at a wavelength λ = 5,000 Å. The results of the computation are shown in the curves of figure III-13.

For satisfaction of the condition $F_0 \gg \nu_\Sigma f_0$ (for example, by a factor of 10) it is necessary to have $\nu_\Sigma = 2{,}6 \cdot 10^{-8}$. Consequently, a molecular oscillator must be used as the frequency standard on the ground.

(b) $F_0 \ll \nu_\Sigma f_0$ with $\lambda^*_2 = \lambda > \lambda^*_1$.

Calculations show that even in the case $\Delta f = \Delta f_{d\ min}$ ($\Omega t = 0.\pi$) the condition $F_0 \ll \nu_\Sigma f_0$ is satisfied for all radio links even with the transmission of a wide-band television signal (F_0 = 94 Mcps).

Let us carry out the calculation (similar computations have been made in reference 121) for the required radiation power of the onboard transmitter with the transmission of the television signal (F_0 = 94 Mcps, N_0 = 16, m = 8.4) for two cases:

(1) the ground receiver bandpass width at the radio frequency is constant and equal to the maximal frequency shift due to the Doppler effect;

(2) the ground receiver bandpass width is variable and equal to the instantaneous frequency shift due to the Doppler effect.

We determine P_{t_1} and P_{t_2} for both cases and present the results in the curves of figures III-14 and III-15. /260

(c) $N_m \frac{2F_m}{F} \ll 1$ with $\lambda^*_2 = \lambda > \lambda^*_1$.

Let us make an analysis of the required radiation power of the onboard transmitter with the transmission of a telegraph signal, using amplitude keying (F_m = 5 cps, N_m = 10) if $\sigma = 6{,}2 \cdot 10^{-9}$, $T_0 = 7{,}2 \cdot 10^4$ °K, $\lambda = 5000$ Å, D_2 = 5 m, D_1 = 1 m.

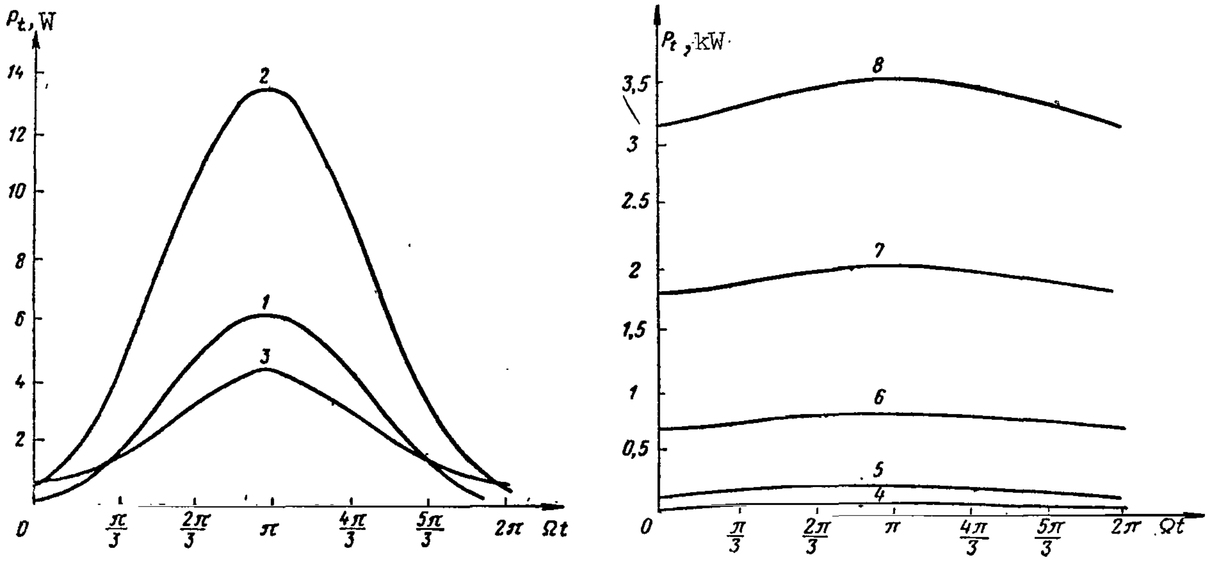

Figure III-13. Variation of radiated power of onboard laser with time for FM television transmission, $F_0 \gg \nu_\Sigma f_0$.

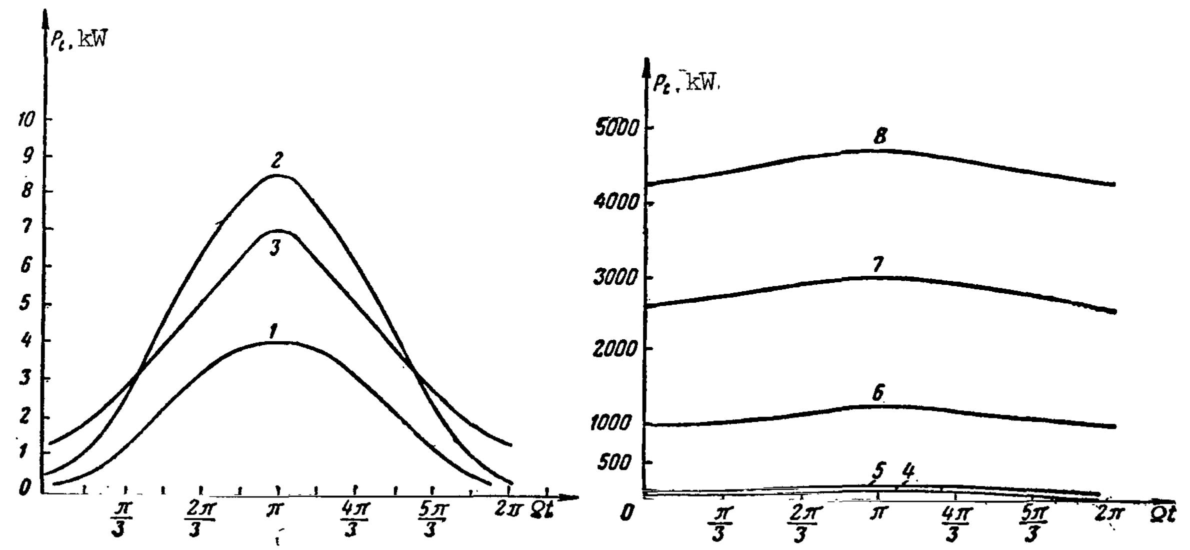

Figure III-14. Variation of radiated power of onboard laser with time for FM television transmission, $F_0 \ll \nu_\Sigma f_0$, $F = \Delta f_{D_{max}}$.

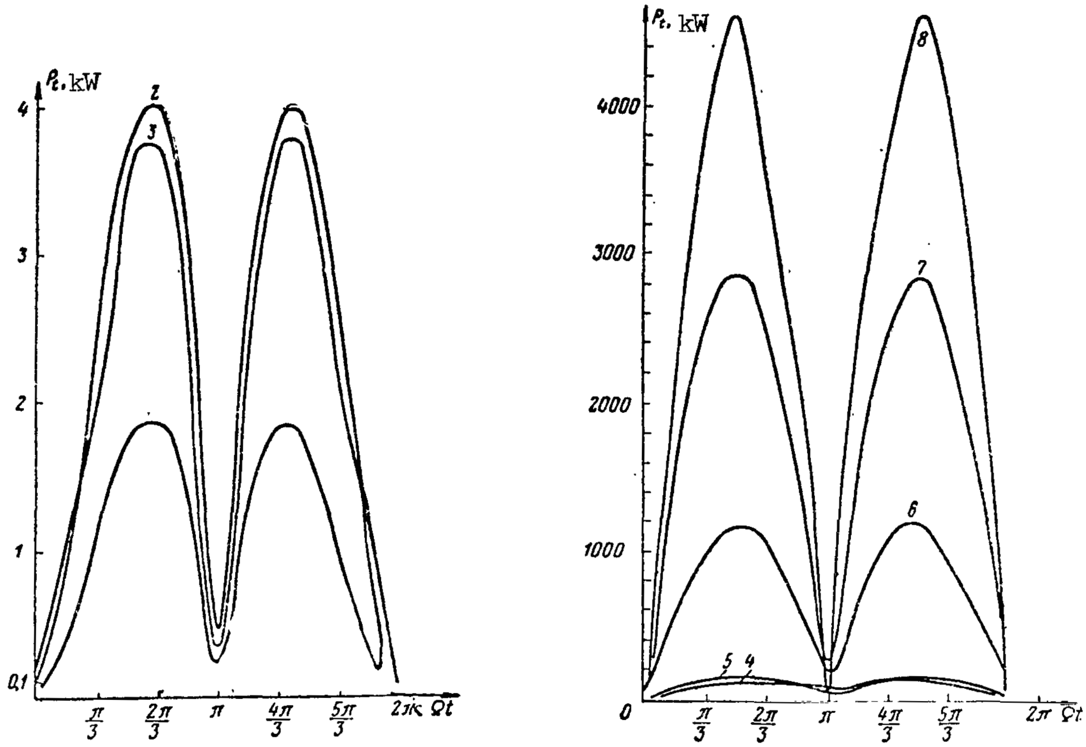

Figure III-15. Variation of radiated power of onboard laser with time for FM television transmission, $F_0 \ll \nu_\Sigma f_0$, $F = \Delta f_D(t)$.

Let us consider the same two cases:

(1) the passband width of the receiver at the radio frequency is constant and equal to the maximal frequency shift due to the Doppler effect;

(2) the passband width of the receiver at the radio frequency is variable and equal to the instantaneous value of the frequency shift due to the Doppler effect.

We compute P_{t_1} and P_{t_2} and present the results of the computation in the curves of figures III-16 and III-17.

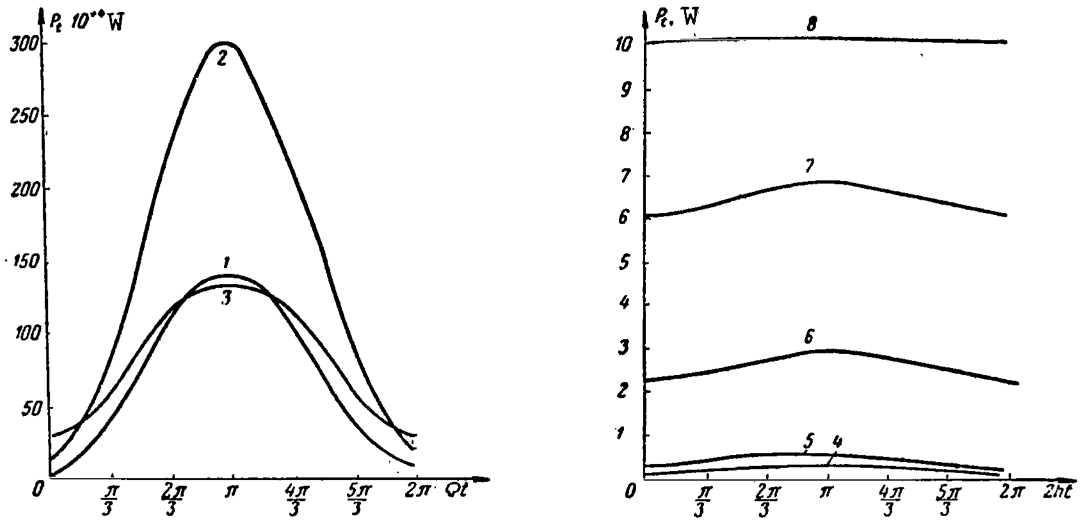

Figure III-16. Variation of radiated power of onboard laser with time for AT telegraphy transmission, $N_m \frac{2F_m}{F} \ll 1$, $F = \Delta f_{D_{max}}$.

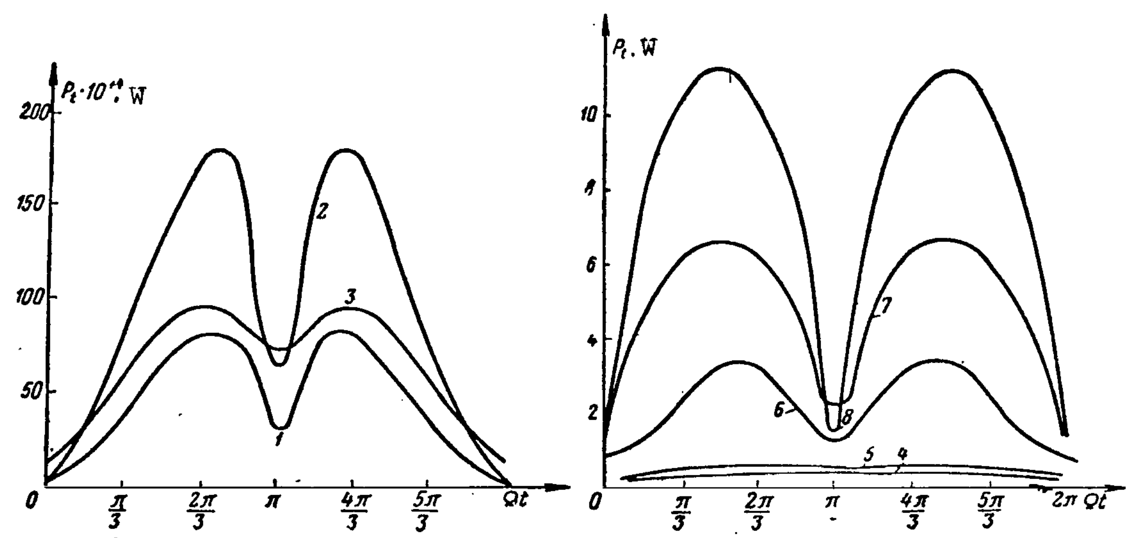

Figure III-17. Variation of radiated power of onboard maser with time for AT telegraphy transmission, $N_m \frac{2F_m}{F} \ll 1$, $F = \Delta f_D(t)$.

5. Communication Possibilities Within the Solar System

The possibilities of maintenance of two-way continuous radio communication with a spacecraft located in the vicinity of or on the surface of various planets of the solar system are limited as the result of the limitations of the capabilities of the various components of the radio link-power and efficiency of the onboard high frequency generators, the geometric dimensions and the accuracy of fabrication of the onboard antennas, the capacity of the power supplies, etc. Assuming we are given the efficiency of the generator (ξ = 10 percent), the geometrical dimensions and the accuracy of the fabrication of the antennas, we shall make an analysis of the possibilities of communications within the limits of the solar system only from the point of view of the possibilities of long-life power sources for the onboard equipment.

The previous consideration of the various power sources permits us to draw the conclusion that the only long-life power source for the onboard equipment at the present stage of development are the solar cells, although this power source is not as promising as, for example, atomic or nuclear energy /261 sources.

Assuming the efficiency of a battery of solar cells ξ' = 10 percent, let us determine the required panel area of a battery of solar cells in the most severe cases (maximal length of the radio link, maximal Doppler frequency shift) for all radio links for which we calculated the power requirements in the preceding section.

Let us make an estimate of the magnitude of the power which can be taken from a panel of solar cells of area $S_1 = 1\ m^2$ with the solar battery located on the surface of the various planets (panel pointed at the Sun). The results of the computations are summarized in table III-9.

TABLE III-9

Planet	Power from solar cell panel with area $S_1 = 1\ m^2$, w
Venus	193
Mars	43.5
Mercury	670
Jupiter	3.7
Saturn	1.1
Uranus	0.28
Neptune	0.11
Pluto	0.064

Let us make an estimate of the required direct current power P_0 on board the spacecraft for the cases $R_{av} = R_{av\ max}$ and $\Delta f_d = \Delta f_{d\ max}$ (assuming that the generator efficiency ξ = 10 percent), and also an estimate of the required area of the panel of solar elements S for these same cases.

Let us find P_0 and S with an omnidirectional antenna on board the spacecraft using FT, F_m = 5 cps, N_m = 10,

$$N_m \frac{2F_m}{F} > 1 \qquad (*)$$

D_{21} = 25 m, D_{22} = 100 m; the results of the computation are summarized in tables III-10 (D_{21} = 25), III-11 (D_{22} = 100 m).

Analysis of the data of tables III-10 and III-11 permits us to draw the following conclusions.

1. With a ground antenna with diameter D_{21} = 25 m, a nondirectional antenna on the spacecraft and compensation of the frequency shifts due to the instabilities and the Doppler effect with an accuracy to tens /262
of cycles per second, constant two-way telegraphic communication with the spacecraft from Earth can be maintained (5 baud speed of operation), if the spacecraft is in the vicinity of or on the surface of Mercury, Venus or Mars.

2. With other conditions remaining equal, but with increase of the diameter of the ground antenna to D_{22} = 100 m, communications can be maintained also in a limited number of cases--with the spacecraft in the vicinity of or on the surface of Mercury, Venus, Mars and, possibly, Jupiter. If we do not take steps to compensate the frequency shifts due to the instabilities and the Doppler shift effect, then, with an omnidirectional antenna on board the spacecraft and with use of frequency keying, two-way telegraphic communication cannot be maintained, even with the spacecraft in the vicinity /263
of or on the surface of the planets closest to Earth.

Changeover of the channel to pulse operation with high duty factor M ($M \approx$ 10,000) changes the situation, since the case (*) is realized and, consequently, the two preceding conclusions will be valid (with the condition that the average transmitter power remains constant).

We determine P_0 and S with a highly directive antenna on board the spacecraft, television transmission, $\sigma_1 = 10^{-4}$, $\sigma_2 = 10^{-5}$, $D_{1,2} = 5$ м, $F_0 \gg \nu_\Sigma f_0$ for $\lambda^*_2 = \lambda > \lambda^*_2$

and summarize the results of the calculations in tables III-12 and III-13 (for σ_1 and σ_2, respectively).

TABLE III-10

Planet	Required maximal direct current power P_0, kw	Required area of solar cell panel, m^2
Venus	2.5	13
Mars	4.8	110
Mercury	1.7	2.5
Jupiter	31.6	8,550
Saturn	80.5	73,000
Uranus	294	105,000
Neptune	691	6,300,000
Pluto	1,200	18,750,000

TABLE III-11

Planet	Required maximal direct current power P_0, w	Required area of solar cell panel, m^2
Venus	184	0.95
Mars	326	7.5
Mercury	126	0.19
Jupiter	2,600	700
Saturn	5,800	5,250
Uranus	22,000	78,600
Neptune	52,100	437,000
Pluto	90,000	1,400,000

TABLE III-12

Planet	Required maximal direct current power P_0, mW	Required area of solar cell panel S, m^2
Venus	$3.5 \cdot 10^{-2}$	1,800
Mars	1.5	24,500
Mercury	$3.95 \cdot 10^{-1}$	590
Jupiter	7.7	2,050,000
Saturn	17.4	16,000,000
Uranus	64	228,000,000
Neptune	150	1,360,000,000
Pluto	258	4,050,000,000

TABLE III-13

Planet	Required maximal direct current power P_0, mW	Required area of solar cell panel S, m^2
Venus	$5.8 \cdot 10^{-4}$	29.5
Mars	$1.0 \cdot 10^{-2}$	245
Mercury	$5.4 \cdot 10^{-3}$	8
Jupiter	$2.75 \cdot 10^{-1}$	73,000
Saturn	$1.75 \cdot 10^{-1}$	160,000
Uranus	$6.4 \cdot 10^{-1}$	2,280,000
Neptune	1.5	33,600,000
Pluto	2.6	40,500,000

Analysis of the data of tables III-12 and III-13 permits us to make the following conclusions:

1. With antennas of diameters $D_{1,2} = 5$ m on the ground and on board the spacecraft and with an accuracy of manufacture of $\sigma_1 = 10^{-4}$, the /264 transmission of high-fidelity television images from on board the spacecraft is not possible at the present stage of development if the craft is on the surface of any planet of the solar system, with the possible exception of Mercury.

2. With antennas of diameter $D_{1,2} = 5$ m on the ground and on board the spacecraft and with fabrication accuracy of $\sigma_2 = 10^{-5}$, transmission of high-fidelity television images from on board the spacecraft at the present stage of development is possible only if the craft is in the vicinity or on the surface of the planets closest to Earth--Mercury, Venus and, possibly, Mars.

We calculate P_0 and S for the case $F_0 >> \nu_\Sigma f_0$, with telephony transmission $\sigma_1 = 10^{-4}$, $\sigma_2 = 10^{-5}$, $\lambda^*_2 = \lambda > \lambda^*_1$, $D_{1,2} = 5$ м and summarize the results in tables III-14 and III-15.

Analysis of the data of tables III-14, III-15 permits us to draw /265 the following conclusions:

1. With antennas of diameter $D_1 = D_2 = 5$ m on the ground and on board the spacecraft and with a fabrication accuracy of $\sigma_1 = 10^{-4}$, speech transmission from on board the spacecraft is possible with it in the vicinity or on the surface of the planets closest to Earth--Mercury, Venus and Mars.

TABLE III-14

Planet	Required maximal direct current power, w	Required area of solar cell panel, m^2
Venus	250	1.3
Mars	740	17.1
Mercury	280	0.4
Jupiter	5,450	1,460
Saturn	12,400	11,700
Uranus	45,500	162,000
Neptune	108,000	1,980,000
Pluto	165,000	2,900,000

TABLE III-15

Planet	Required maximal direct current power, w	Required area of solar cell panel, m^2
Venus	4.1	0.02
Mars	7.4	0.17
Mercury	3.8	0.005
Jupiter	195	52
Saturn	124	112
Uranus	455	1,620
Neptune	1,080	9,800
Pluto	1,650	29,000

2. With antennas of diameter $D_1 = D_2 = 5$ m on the ground and on board the spacecraft and with a manufacturing accuracy of $\sigma_1 = 10^{-5}$, speech transmission from on board the spacecraft is possible with it on the surface or in the vicinity of Mercury, Venus, Mars, Jupiter and Saturn.

In both cases a molecular oscillator should be selected as the frequency standard on the ground. The values of P_0 and S calculated for the case $G_1 = G_{1\,max}$ FT transmission with $F_m = 5$ cps, $N_m = 10$ and $N_m \frac{2F_m}{F} \ll 1$ with $\lambda^*_2 = \lambda > \lambda^*_1$, $D_2 = 100$ m, $D_1 = 5$ m are summarized in table III-16.

TABLE III-16

Planet	Required maximal direct current power, w		Required area of solar cell panel, m^2	
	$\sigma_1 = 10^{-4}$	$\sigma_2 = 10^{-5}$	$\sigma_1 = 10^{-4}$	$\sigma_2 = 10^{-5}$
Venus	23	1.2	0.12	0.006
Mars	38	1.2	0.87	0.027
Mercury	26	1.1	0.038	0.002
Jupiter	477	47.7	138.0	14.0
Saturn	672	21.0	610	19.0
Uranus	3,840	120	13,700	430
Neptune	9,020	280	82,000	2,600
Pluto	15,100	470	236,000	7,400

Analysis of the data of table III-16 permits us to draw the following conclusions:

1. With an antenna on the ground with diameter of $D_2 = 100$ m and a spacecraft onboard antenna with a diameter $D_1 = 5$ m and with a fabrication accuracy of both antennas $\sigma_1 = 10^{-4}$, transmission of telegraphic information from on board the spacecraft without compensation on the ground for the frequency shifts resulting from the instabilities and the Doppler effect is /266 possible with the spacecraft in the vicinity of or on the surface of Mercury, Venus, Mars and, possibly, Jupiter.

2. With the other conditions remaining the same, but with the antenna fabrication accuracy $\sigma_2 = 10^{-5}$, communications are possible with the spacecraft in the vicinity or on the surface of Mercury, Venus, Mars, Jupiter, Saturn and, possibly, Uranus.

With switchover of the channel to the pulse methods of operation (OTM) with high duty factor M, the power expenditures diminish by nearly two orders of magnitude and two-way communication becomes possible with $\sigma_1 = 10^{-4}$, and with the spacecraft in the vicinity or on the surface of Mercury, Venus, Mars, Jupiter, Saturn and Uranus. With $\sigma_2 = {}^{-5}$, and the use of the pulse methods of operation, telegraphic communication of the spacecraft with Earth becomes possible with the spacecraft in the vicinity of or on the surface of any planet of the solar system.

We determine P_0 and S with a laser on board the spacecraft, transmission of television with $F_m = 5$ Mcps, $N_m = 10^4$, $F_0 \gg \nu_\Sigma f_0$, $D_2 = 5$ m, $D_1 = 1$m, $\sigma = 6.2 \cdot 10^{-9}$, and present the results in table III-17.

TABLE III-17

Planet	Required maximal direct current power, w	Required area of solar cell panel, m^2
Venus	6.4	0.33
Mars	138	3.2
Mercury	46	0.07
Jupiter	830	225
Saturn	2,450	2,220
Uranus	8,800	31,500
Neptune	20,800	190,000
Pluto	26,000	565,000

Analysis of the data of table III-17 permits us to conclude that with the laser on board the spacecraft transmission of wideband television images is possible from on board the spacecraft with it on the surface of Mercury, Venus, Mars and Jupiter. Here it is assumed that a molecular oscillator is used as a frequency standard on the ground.

The values of P_0 and S calculated for the case of the laser on board and television transmission with $N_m = 10^4$, $F_m = 5$ Mcps, $F_0 \ll \nu_\Sigma f_0$, $D_2 = 5$ m, $D_1 = 1$ m, $\sigma = 6.2 \cdot 10^{-9}$; $F = \Delta f_{d\ max}$ are summarized in table III-18. /267

TABLE III-18

Planet	Required maximal direct current power, w	Required area of solar cell panel, m^2
Venus	39,800	204
Mars	85,000	1,950
Mercury	72,000	108
Jupiter	1,400,000	380,000
Saturn	1,680,000	1,520,000
Uranus	12,620,000	45,000,000
Neptune	29,800,000	271,000,000
Pluto	47,400,000	740,000,000

Analysis of the data of table III-18 permits us to conclude that with the laser on board the spacecraft any continuous information (television, telephony) from on board can be transmitted, without compensation on the ground for the frequency shift resulting from the Doppler effect, only with the spacecraft and on the surface of the planets closest to Earth--Mercury and Venus.

The values of P_0 and S computed for the case of the laser on board, transmission of AT telegraphy with $F_m = 5$ cps, $N_m = 10$, $N_m \frac{2F_m}{F} \ll 1$, $D_2 = 5$, $D_1 = 1$ m, $\sigma = 6.2 \cdot 10^{-9}$, $F = \Delta f_{d\ max}$, are summarized in table III-19.

TABLE III-19

Planet	Required maximal direct current power, w	Required area of solar cell panel, m^2
Venus	0.15	$7 \cdot 10^{-4}$
Mars	0.30	$7 \cdot 10^{-3}$
Mercury	0.16	$2 \cdot 10^{-4}$
Jupiter	2.96	0.8
Saturn	5.56	5.05
Uranus	29.10	104
Neptune	69.0	626
Pluto	115.5	1,800

Analysis of the data of table III-19 permits us to conclude that /268 with the use on board the spacecraft of a laser with diameter of the telescope of the optical system $D_1 = 1$ m, and a telescope with $D_2 = 5$ m on the ground, two-way telegraphic communication (amplitude keying, $N_m = 10$, $F_m = 5$ cps) without compensation of the Doppler frequency shift can be maintained with the spacecraft in the vicinity of or on the surface of the planets Mercury, Venus, Mars, Jupiter, Saturn and Uranus.

With switchover of the channel with these parameters to pulse operation (PTM) with high duty factor M, the power expenditure is reduced by $\sqrt{M}$ times (refs. 21 and 43), and with $M > 1{,}000$ two-way telegraphic communication can be maintained between Earth and the spacecraft located in the vicinity of or on the surface of any planet of the solar system.

The increased range of communications with the use of laser on board the spacecraft is completely defined by the ground frequency standards. For example, if a ground receiver heterodyne frequency stability $\nu_{inst} = 10^{-10}$ is provided, speech transmission becomes possible (with high fidelity) from the spacecraft when it is located in the vicinity of or on the surface of any planet of the solar system.

CHAPTER IV. EXPERIMENTAL SYSTEMS FOR RADIO COMMUNICATION WITH SATELLITES AND SPACECRAFT

Chapters I and II presented the theoretical analysis of systems /269 for surface communication, using relaying of radio signals via AES and the Moon, and the calculation of the power requirement of the satellite-Earth radio links; they also presented estimates of the technical and economic effectiveness of these communications systems, based on the present level of development of rocketry and electronics. Chapter III studies--again theoretically--the possibilities of maintaining two-way radio communication with spacecraft located in the vicinity (or on the surface) of the various planets of the solar system. There is already extensive experimental data on maintenance of radio communication with various vehicles in space--satellites, luniks, space rockets sent to Mars and Venus, etc. There are also data on experimental systems for surface communication by means of relaying of radio signals via AES.

It is obviously of interest to present some of this material and to see how the problems of communication with spacecraft, which were analyzed theoretically in the preceding chapters, have been resolved in practice.

Since at the present time the U.S. and the USSR have a monopoly in the field of launching of satellites and space rockets, we shall consider separately the communication systems with the space vehicles launched by the USSR and the U.S.

1. Systems for Radio Communication with Space Vehicles Launched in the USSR (ref. 82)

In the present section we shall present only brief information on /270 the systems of radio communication used on the space vehicles launched in the USSR, without a detailed description of the vehicles, their flight trajectories and the scientific problems solved, since this material goes beyond the framework of the present book and has been quite well covered in the literature (ref. 82).

First Artificial Earth Satellite. The first artificial Earth satellite, launched Oct. 4, 1957, was the first space vehicle in the history of man to send signals to Earth. Thus, this was the first space radio link.

The satellite traveled in an elliptical orbit with apogee to 900 km with orbital inclination to the equatorial plane of 65^{o}, and had a spherical shape with diameter of 58 cm and a weight of 83.6 kg (ref. 82). Two radio transmitters were on board, radiating continuously at frequencies of f_1 = 20.005 Mcps

(λ_1 = 15 m) and f_2 = 40.002 Mcps (λ_2 = 7.5 m). Both transmitters used frequency keying and transmitted dots with an average duration of the elementary trains of τ = 0.3 sec.

Four whip antennas were installed on the outer surface of the satellite housing with lengths from 2.4 to 2.9 m. During the injection of the satellite into orbit, the whip antennas were folded against the rocket housing. After separation of the satellite, the antennas were pivoted on hinges and took up a position tangential to the surface of the satellite. The FM signals emitted by the satellite were used not only for direction finding, but also for the transmission of information on the conditions on the satellite (temperature, pressures, etc.). This information was transmitted by means of variation of the duration of the trains radiated by the transmitters (pulse width modulation). Reception of the signals from the satellite was accomplished both at special radio reception centers, and by numerous radio amateurs, and completely confirmed the possibility of reliable radio communication with artificial Earth satellites (ref. 98). Electric power sources located on board /271 the satellite provided for the operation of all the onboard equipment for a period of three weeks.

Second Artificial Earth Satellite. The second AES was launched in the Soviet Union on Nov. 3, 1957. The satellite contained a pressurized cabin with an experimental animal (a dog nicknamed Layka). The weight of the satellite equipment, the experimental animal and the electric power supplies amounted to 508.3 kg (ref. 82).

This satellite used the same transmitters as the first AES, operating at the frequencies f_1 = 20.005 Mcps and f_2 = 40.002 Mcps. The transmitter of the f_1 frequency operated in the same regime as on the first AES, while the transmitter on the f_2 frequency operated intermittently.

The radio signals from the transmitters were used both for the measurement of the orbital parameters of the AES and for the study of the propagation of radio waves.

The second AES traveled along an elliptical orbit with an apogee exceeding 1,700 km with an orbital period of 1 h 42 min, with orbital inclination to the equatorial plane of approximately 65° (ref. 98).

Radio observations of the AES were made at stations located at differing geographic latitudes and longitudes by radar direction finding stations and by DOSAAF clubs. Measurements were made of the field intensity of the signals from the AES, which permitted an evaluation of the degree of absorption of the radio waves in the atmosphere, including those regions which lie above the ionization maximum of the primary ionospheric F_2 layer. (These regions are inaccessible for conventional measurements made from the surface of Earth.)

The measurements showed that at the 15 m wavelength the signals were received far beyond the limits of direct visibility, reaching 10,000-12,000 and even 15,000 km (antipodal effect).

Interesting results were obtained from the observation of the Doppler effect with the aid of recording (on magnetic tape) the variation of the beat tone between the frequency of the radio signals from the AES and the frequency of local stable heterodyne.

Third Artificial Earth Satellite. The third AES was launched on May 15, 1958 in the Soviet Union. The satellite weighed 1,327 kg, and the total weight of the scientific equipment and the power supplies was 968 kg (refs. 82 and 98). /272

The satellite shape was nearly conical with a length of 3.56 m and a maximal diameter of 1.73 m (without accounting for the extended antennas).

The satellite was launched into an elliptical orbit with apogee of 1,880 km. The period of revolution in the beginning was 105.95 min. The number of orbits per day was about fourteen. The satellite was equipped with advanced radio technical equipment, which made it possible to conduct accurate measurements of its movement along the orbit, and with radio telemetric equipment permitting continuous recording of the results of the scientific measurements, the storage of these measurements in the course of the travel of the satellite and transmission to Earth, as the satellite flew over special stations located on the territory of the USSR to receive the stored information.

All instrumentation, scientific and radio technical equipment was fabricated with extensive use of the new semiconductor elements. (The total number of such elements on board the AES amounted to several thousand.) Electric power supply was provided by electrochemical current sources and by silicon semiconductor batteries, which converted solar ray energy into electrical energy.

The Mayak radio transmitter installed on board the satellite continuously radiated signals in the form of telegraphic trains at a frequency of 20.005 Mcps. The radiation power of 250 mW provided for reception at ranges amounting to several thousand kilometers. Telemetric information was transmitted over the radio channel of the Mayak transmitter.

The satellite entered the dense layers of the atmosphere on April 6, 1960 and ceased to exist.

First Space Rocket to the Moon. The launches of the first three AES attained the first cosmic (orbital) velocity of 8 km per sec.

The successes of the Soviet Union in the field of rocketry made it possible to attain the second cosmic (escape) velocity of 11.2 km per sec, which makes interplanetary flight possible. This made possible the launch on Jan. 2, 1959 of the space rocket to the Moon. The weight of the last stage of the rocket was 1,472 kg (without fuel). The following equipment was installed /273 on board the last stage of the rocket for flight observation (ref. 82):

--a radio transmitter radiating telegraphic trains of 0.8 and 1.6 sec duration at frequencies of 19.997 and 19.995 Mcps;

--a radio transmitter operating on the frequency of 19.993 Mcps with telegraphic trains of variable duration of the order of 0.5-0.9 sec, used to transmit data from the scientific observations;

--a radio transmitter radiating at a frequency of 183.6 Mcps, used for measurement of the motion parameters and for transmission of scientific information to Earth.

The flight of the rocket to the Moon lasted 34 hours. At the time of closest approach of the rocket to the Moon the distance between them was 5,000-6,000 km, i.e., approximately one and a half times the Moon's diameter.

On approximately Jan. 7-8, the Soviet space rocket went into an independent orbit about the Sun, becoming its satellite and the first artificial planet of the solar system in history (ref. 91).

A large complex of instrumentation located over the entire territory of the Soviet Union was used for observation of the flight of the space rocket, for measurement of its orbital parameters and for reception of the data of the scientific measurements from on board.

This instrumentation complex included the following radio facilities: a group of automated radars for precise determination of the elements of the initial portion of the orbit; radio telemetry stations for recording of the scientific information transmitted from on board the rocket; the radio technical system for the control of the elements of the trajectory of the rocket at great distances from Earth, and radio technical stations used for the reception of the signals at the frequencies of 19.997, 19.995 and 19.993 Mcps.

The telemetry information transmitted from on board the rocket was recorded at the ground stations on magnetic tapes and photo tapes. To provide for greater reception range, use was made of high-sensitivity receivers and special antennas with large effective area.

Control of the rocket trajectory up to distances of 400,000- /274
500,000 km was accomplished with the aid of a radio technical system operating at a frequency of 183.6 Mcps (ref. 91).

Second Space Rocket to the Moon. On Sept. 12, 1959 the Soviet Union launched the second space rocket to the Moon. The last stage of the space rocket was a controllable rocket weighing 1,511 kg (without fuel), and containing a compartment in the form of a sphere with scientific equipment with a total weight of 390.2 kg (ref. 82).

The following equipment was installed on board the rocket:

--a radio transmitter operating on frequencies of 20.003 and 19.997 Mcps;

--a radio transmitter operating on frequencies of 19.993 and 39.986 Mcps;

--a radio transmitter operating on the 183.6 Mcps frequency.

On Sept. 14, 1959 the Soviet space rocket reached the surface of the Moon. It is difficult to overestimate the importance of this event: for the first time in the history of man a space flight was accomplished from Earth to another heavenly body.

Operation of the radio equipment installed in the container with the scientific and measuring apparatus stopped at the instant of impact of the rocket on the Moon. The successful operation of the radio equipment made possible continuous monitoring of the actual flight trajectory in comparison with the calculated data, permitted reliable prognostication of impact on the Moon and made it possible to determine the region of the impact (to the east of the Sea of Serenity).

Third Space Rocket to the Moon. Photographing of the Back Side of the Moon. Oct. 4, 1959 a multistage booster rocket was used to launch an automatic interplanetary station (AIS) toward the Moon. The most important task of the station was to obtain a photographic image of the back side of the Moon and to transmit this image to Earth (refs. 82 and 99).

The AIS was cylindrical with spherical end closures and contained onboard equipment and chemical power supplies. A portion of the scientific /275
instruments, the antennas and the solar battery sections were located on the outside. The maximal cross-sectional dimension of the station was 1,200 mm and the length was 1,300 mm.

The radio technical equipment on board the station provided for the measurement of the orbital parameters of the station, transmission to Earth of the television and telemetric information, and also the reception of the commands from Earth for the control of the onboard equipment.

All the control of operation of the onboard equipment of the station was accomplished from ground stations using the radio link, and also by autonomous onboard programming devices. Such a combined system provided for the most convenient control of the conduct of the scientific experiments, obtaining information from any regions of the trajectory within the limits of the radio visibility from the ground observation stations.

The electric power supply system contained autonomous units of chemical current sources which supplied the equipment, which operated only briefly, and also a centralized block consisting of a buffer chemical battery. Compensation for the expended energy of the buffer battery was accomplished by solar current sources. The supply of the onboard equipment was provided by means of converting and stabilizing devices.

The process of the photography of the lunar surface and the processing of the film was accomplished automatically according to a preset program (ref. 99). On termination of photographing, the film entered a small automatic processing device, where it was developed, fixed and dried. After this the film entered a special cassette for the transmission of the image.

The transmission of the images of the Moon was accomplished on command from the ground. These commands energized the power supply for the onboard television equipment, the film advance, and the connection of the television equipment to the onboard transmitter. For the extremely long distance transmission with very low transmitter power (several watts), use was made of a rate of transmission of the image which was tens of thousands of times slower than the rate of transmission of the conventional broadcast television centers (ref. 99). Provision was made for transmission of the images in two regimes: a slower transmission at the greater distances and a faster transmission at the shorter distances, during the return flight toward Earth. /276 The number of lines into which the image was broken down could be varied as a function of the selected transmission regime. The maximal number of lines was 1,000 per frame (ref. 99). For synchronizing transmitting and receiving scanning devices, use was made of a method which ensured high noise immunity and high reliability of the operation of the equipment.

The radio link provided for two-way transmission of radio signals. The uplink carried the command signals which controlled the operation of the onboard equipment. The downlink carried the television signals, the signals with the indications of the scientific instruments and the signals for the measurement of the parameters of the motion of the station itself. The ground equipment included powerful radio transmitters, high-sensitivity receivers and recording equipment, and also the receiving and transmitting antenna equipment. The onboard radio equipment of the automatic interplanetary station included the transmitting, receiving and antenna equipment, and also the command and programming radio technical devices.

The images of the Moon were transmitted from on board the AIS over the radio communication line, which at the same time served for the measurement of the parameters of the motion of the station itself.

This combination of functions in a single radio communication line with continuous radiation was accomplished for the first time, and made it possible to provide reliable radio communication up to the maximal distances with the least energy consumption on board.

All equipment of the radio communication link, both on the ground and on board, was duplicated in order to improve the communications reliability. In the case of failure of one of the radio technical instruments on board or the end of its programmed service life, it could be replaced by the reserve instrument by transmission of the proper command from the ground control station.

The onboard radio equipment utilized semiconductors, ferrites and other advanced components and materials. Special attention was devoted to achieving minimal weight and volume of the instruments, which made it possible to increase the weight and volume available for the electric power supplies. From considerations of conservation of electrical energy, the power radiated by the onboard radio transmitters was limited to a few watts.

The power for the onboard equipment of the interplanetary station was supplied from self-contained units of chemical power supply sources and from a centralized power supply system. This system included a solar battery, whose individual sections were located on the external surface of the spacecraft, and a chemical buffer battery. The consumption of the energy of the buffer battery during operation of the onboard equipment was compensated by the energy coming from the solar battery. The onboard equipment was supplied through transforming and stabilizing devices.

In order that communication with the station was not interrupted as it rotated, the station antennas radiated radio signals uniformly in all directions.

The signal received on Earth was very weak. At the maximal distance of the station from Earth, the received portion of the power radiated by the onboard transmitter was 100 million times less than the average power received by a conventional television receiver. Reception of such weak signals could be accomplished only by very sensitive receiving devices with a low level of self-noise.

The economical utilization of the power supplies on board the station, the use of a radio communication link with continuous radiation and combined functions, the use on the ground of special receiving antennas, high-sensitivity receiving equipment, the use of special methods for the processing and the transmission of the signals--all this made it possible to ensure reliable radio communication with the AIS, trouble-free operation of the radio command link and the accomplishment of the planned reception of the images of the Moon and the telemetric information.

The reception on Earth of the signals of the Moon images was accomplished on special devices for the recording of the television images on photographic film: on magnetic recording equipment with high stability of the magnetic tape travel speed, on skiatrons (cathode ray tubes with long-term storage of the image on the screen), and on equipment for visual recording with registration of the image on electrothermal paper. The data obtained from all forms of recordings were used for the study of the back /278 side of the Moon.

The radio and television equipment on board the AIS provided for transmission of the images up to a distance of 470,000 km (refs. 82 and 99).

This operation gave experimental confirmation of the possibility of transmission in outer space at extremely great distances of half-tone images of high clarity without significant specific distortions in the process of the propagation of the radio signals.

Launch of the First Spacecraft-Satellite with Experimental Animals. The launch of a large spacecraft into an Earth satellite orbit was accomplished on Aug. 19, 1960. The cabin, equipped with everything necessary for the future flight of man, contained experimental animals, including the dogs named Strelka and Belka. The satellite spacecraft was injected into a nearly circular orbit with an altitude of about 300 km; the weight of the spacecraft without the last stage of the booster rocket was 600 kg (ref. 82).

The spacecraft carried radio telemetry equipment for transmission to Earth of data on the status of the experimental animals and the operation of all the systems installed on board.

This equipment operated in two ways:

(a) direct transmission of the telemetry information;

(b) memorizing (storage) of the information, with its later transmission during flight over the receiving stations.

The power supply for the onboard equipment was provided from chemical current sources and from a solar battery. The solar battery was located on two semidisks of 1,000 mm diameter oriented to the Sun by a special system independently of the attitude of the spacecraft.

Special television equipment was installed on board for the transmission of the image of the experimental animals. This equipment included two miniature television cameras, one transmitting a front view of Belka and the other a /279 side view of Strelka. The television transmission began before liftoff of the spacecraft, and the condition of the animals was observed during liftoff, at the time of transition from accelerated flight to weightlessness and in orbit. The cameras were turned on and off by the radio link from Earth.

During the 18th orbit of the ship the descent command was sent over the radio link. The systems for the control and deceleration installed on board the satellite provided for descent of the ship at the designated point with an accuracy of about 10 km.

Thus, for the first time, living creatures had completed a space flight and successfully returned to Earth. In addition, the transmission of images of moving objects was accomplished for the first time.

Launch of a Space Rocket to the Planet Venus. On Feb. 12, 1961 in the Soviet Union an advanced multistage rocket launched a heavy AES into a circular orbit about Earth (with $h \approx 200$ km). That same day a controlled space rocket was launched from this satellite, which injected an AIS into a trajectory to the planet Venus (ref. 82).

An automated electronic instrumentation complex was developed for the control of the AIS, determination of its orbit and for two-way communication with the AIS.

The development of the complex faced Soviet scientists and engineers with a number of serious problems associated with the provision for communication over tremendous distances, with rigid requirements on the accuracy of the determination of the coordinates and on the reliability of the operation of the equipment in the course of a long period of time.

The entire trajectory of the flight of the space rocket can be divided arbitrarily into three segments: the segment of the flight of the heavy artificial Earth satellite, the segment of the launch of the space rocket from the heavy satellite and the segment of the motion of the AIS under the influence of gravitational forces in the direction of Venus.

The measurement of the elements of the trajectory of the heavy satellite was accomplished by special equipment located on the territory of the Soviet Union. Information on the operation of the components and systems /280
of the satellite was received by the radio telemetry stations located both in the Soviet Union and on board special ships in the oceans. The launch of the space rocket from the heavy satellite was monitored by the telemetry systems.

After separation of the AIS, the ground-based instrumentation complex intended for the conduct of the orbital and telemetric measurements came into operation. At each instrumentation station of the ground-based network there were installed special radio and electronic transmitting and receiving recording devices, parabolic antennas with equipment for programmed tracking.

The determination of the actual orbit, when the AIS was at a distance from Earth greater than 100,000 km, was accomplished by the electronic equipment of the Center for Deep Space Radio Communication, which provided for reception of the telemetry information and the control of the equipment on board the AIS throughout the flight. The radio command link was used to turn on and off the corresponding instruments of the AIS, to change the rate of transmission of the telemetry information, to switch the power supply sources, etc.

The operation of all equipment during the most remote portion of the flight of the AIS was controlled by a special program, which determined the duration of the communication periods, the intervals between these periods and the operating regime of the equipment.

For the reception of the radio signals at extremely great distances use was made of narrow-band, low-noise receiving equipment, which required precise computation of the frequency shift due to the Doppler effect.

With the interplanetary station at distances measured in tens and hundreds of millions of kilometers, the signal power reaching Earth is very small. Thus, for example, at a distance of 70 million km only 10^{-22} W arrives on one square meter of Earth's surface. For the reception of these very weak signals large antenna structures were built in the Center for Deep Space Radio Communication, to track the AIS automatically in accordance with a program.

The antenna could be pointed to any point of the celestial sphere with an accuracy of a few angular minutes. The tracking programs were automatically entered into an electronic computer which controlled the antennas.

Four antennas were installed on the outside of the spacecraft. /281
One of them--highly directional--had the shape of a paraboloid with a diameter of about 2 m; it was intended for communication with the interplanetary station at great distances from Earth and transmission of a large volume of information in the course of a short period of time. The two cruciform antennas installed on the solar battery panel were nondirectional and were intended for communication at moderate distances from Earth. An omnidirectional antenna--a whip 2.4 m long--was intended for the transmission of information and the determination of the trajectory parameters in the near-Earth segment.

The maximum dimensions of the station (without account for the antennas and solar batteries) were 2,035 mm in length and 1,050 mm in diameter.

The weight of the AIS was 643.5 kg.

Until the separation of the station from the space rocket, the solar battery panels and the cruciform and whip antennas were folded and were extended immediately after separation. The parabolic antenna was to be opened on approach to Venus.

The electronic complex of the AIS resolved the following problems:

--measurement of the parameters of the motion of the station relative to Earth;

--transmission to Earth of the results of the measurements made by the scientific equipment on board;

--transmission to Earth of information on the operation of the onboard instruments, pressure and temperature inside the vehicle and skin temperature;

--reception from Earth of the radio commands for the control of the operation of the onboard equipment.

The control of the operation of the onboard equipment of the station was accomplished by means of transmission of commands over the radio link from the ground stations and also by self-contained onboard programming devices.

Periods of communication on both uplinks and downlinks were accomplished successfully up to distances of several million kilometers; after this communications were terminated for reasons not known. Thus, for the first time the launch of a controlled vehicle from an artificial satellite of /282

Earth into an interplanetary trajectory was accomplished as well as control of the vehicle at distances of several million kilometers (ref. 82).

The First "Vostok" Piloted Spacecraft. On Apr. 12, 1961 in the Soviet Union, flight of man into cosmic space was accomplished for the first time in history. The Vostok spacecraft with pilot-cosmonaut Yu.A. Gagarin of the USSR on board was injected into orbit as a satellite of Earth. The weight of the spacecraft without the last stage of the booster rocket was 4,725 kg. The orbital perigee altitude was 181 km, the apogee altitude was 327 km and the orbital inclination was 64°57' (ref. 82).

A multitude of radiometric and radiotelemetric devices was installed on board the spacecraft for the measurement of the orbital parameters and for the monitoring of the operation of the onboard equipment. The measurements of the parameters of the motion of the spacecraft and the reception of the telemetric information during flight were performed by ground-based stations located on the territory of the USSR. The measurement data were transmitted automatically over communication lines to the computer centers, where they were processed on electronic computers. Thus, in the course of the flight, operational information was obtained on the principal orbital parameters, and the future path of the ship was predicted.

There was also a signal radio system on board the spacecraft operating on a frequency of 19.995 Mcps. This system served for tracking the ship and for transmission of a portion of the telemetric information.

The television system transmitted to the ground an image of the cosmonaut to provide visual monitoring of his condition. One of the television cameras presented a face view and the other a side view of the pilot.

Two-way communication of the cosmonaut with Earth was provided by a radiotelephone system operating in the shortwave (19.019 and 20.006 Mcps) and ultrashortwave (143.625 Mcps) bands.

The UHF channel was used for communication with ground stations at /283
distances up to 1,500-2,000 km. Experience showed that communication using the HF channel with ground stations located on the territory of the USSR could be provided over most of the orbit.

The radiotelephone system included a magnetic recorder which permitted recording of the cosmonaut's speech in flight, with later reproduction and transmission during flight of the spacecraft over the ground receiving stations. Provision was also made for radiotelegraphic transmission from the spacecraft.

The first flight of man into cosmic space in the history of mankind, accomplished by the Soviet cosmonaut Yu.A. Gagarin on the Vostok spacecraft, provided a tremendous amount of scientific information on the practical possibility of manned flight into space and opened up a new--cosmic--era in the history of man.

After the flight of Vostok-1 came the flight of Vostok-2 with pilot-cosmonaut Major G.S. Titov on board; the dual flight of P.R. Popovich and G.A.G. Nikolayev (Vostok-3 and Vostok-4), the flight of V.F. Bykovskiy and the first woman-cosmonaut V.V. Nikolayeva-Tereshkova (Vostok-5 and Vostok-6).

Consistent telephonic and telegraphic communication with all cosmonauts was provided on the HF and UHF bands, and regular transmission of television images of the cosmonauts and the interior of the spacecraft cabin from space was accomplished. The quality of the images transmitted was considerably better than the quality of the images of the astronauts transmitted from on board the inhabited spacecraft of the U.S. (flights of Cooper, Glenn and others).

Launch of the Space Rocket to the Planet Mars. On Nov. 1, 1962 the AIS Mars-1 was launched in the Soviet Union toward the planet Mars.

The maximal dimensions of the station were: 3,300 mm long, 1,100 mm diameter of the orbital compartment and 4,000 mm diameter with account for the solar batteries and the radiators. The station weighed 893.5 kg.

The station was equipped with radio technical apparatus, a system for /284
orientation and correction of the trajectory, electric power supply sources. The onboard radio systems were used for the trajectory measurements and the transmission of the telemetry information to Earth.

Three radio systems were installed on board the station, operating in the metric (1.6 m), decimetric (32 cm) and centimetric (5 and 8 cm) bands. For the transmission of telemetry information there were several commutators installed on board the station which were connected in sequence to the scientific apparatus transmitter and to the sensors recording the status of the station during the time of radio communication with Earth. The memory device recorded the indications of the scientific instruments during the periods between communications and transmitted the information to Earth during the communication periods.

In addition to the transmission of the information on the status of the station, the radio complex operating in the metric wavelength band served for maintaining communication with Earth in the case of abnormal operation of the orientation system.

The periods of communication with the station were controlled both automatically and on command from Earth. The program for the operation of the onboard systems provided for automatic communication periods with intervals of two, five and fifteen days. Selection of the interval between the communication periods was made using the radio command link from Earth. The intervals between communication periods were required, first, in order that the solar batteries could charge the buffer chemical battery whose energy is consumed during the communication periods and, second, in order that the communication take place at a time of best radio visibility of the station, which repeats every day as a result of the rotation of Earth about its axis.

Prior to Dec. 13, 1962, the station operated in the mode with regular contact every other day; then the communication contacts were made every five days. Each contact began with the reception of telemetry information containing the results of scientific measurements and data on the conditions in the station, the measurements of its velocity and range. Then, on command from Earth, the memory device was activated for the reproduction of the /285 previously recorded information. The contact terminated with the reception of the telemetry information on the status of the station at the termination of the operation.

By Feb. 14, 1963, the interplanetary station had reached a distance of about 50 million km from Earth. During this time a large number of radio contacts had been made, and valuable scientific information had been transmitted over the radio telemetry link.

The maximal range of communication with the Mars-1 space rocket was about 100 million km. Thus, the Soviet Union was in the lead not only in the development of near-Earth space, but also in the mastery of interplanetary (deep) space. Communication with the U.S. space rocket Mariner was maintained only to a distance of 80 million km (refs. 100 and 101).

2. Systems for Communication with Space Vehicles Launched in the U.S.

The first AES was launched by the U.S. in 1958--a year after the first launching of an AES in the USSR. Since that time work has gone forward both in the direction of the study of "near" space (study of the radiation belts, cosmic rays, meteor fluxes, Earth's magnetic field, man in space, etc.) and the study of "deep" space (the study of the properties of the interplanetary medium, the study of the planets of the solar system, etc.). Experiments are also being conducted on the study of the possibilities of the use of AES as relay stations (active and passive) for surface communications systems. We shall dwell in somewhat greater detail on this last question.

Work on the creation of the first communications satellites has been carried out in the U.S. by several firms under the direction of NASA and ARPA. At the present time firms of other foreign countries (England, France, West Germany, Italy, Japan, Brazil) are also participating in the studies of /286 the creation of experimental systems for surface communications by means of relaying via AES.

Experiments have been conducted on both passive and active relaying of radio signals via AES.

The first active radio repeater, Score, was launched on Dec. 18, 1958 from Cape Canaveral, using an Atlas booster into an inclined ($i = 32.3^{o}$) elliptical orbit with an apogee $h_a = 1{,}481$ km and a perigee $h_p = 177$ km.

The satellite consisted of the Atlas booster rocket itself, with length 25.7 m and diameter 3.04 m. The satellite equipment consisted of 2 transceivers operating at frequencies of 132.435 and 132.095 Mcps. Operation was in the delayed relay mode. Storage of the signal sent by the ground transmitting station was accomplished by recording on magnetic tape (ref. 102).

The Score AES receiver was a modification of a commercial FM receiver of the superheterodyne type using semiconductors. The frequency band was UHF with a nominal value of 150 Mcps. The receiver sensitivity was 2 μV with an IF amplifier bandpass width of 40 kcps. The receiver weighed 340 g and the power consumption was 2.5 W.

The onboard transmitter was a vacuum-tube FM transmitter with frequency deviation of 5 kcps, radiation power of 8 W and power consumption of 39 W. The power supply consisted of 18 V, 45 A-hr silver-zinc batteries.

Reception and transmission were accomplished on two separated antennas of the slot type, having an approximately isotropic pattern; the satellite was not stabilized (ref. 102).

Transmission and reception of the information was accomplished by four ground stations equipped with transmitting and receiving cophased antennas of the quadspiral type, with diameter of each spiral 0.75 m, length 3.6 m and gain of 9-16 db with respect to an isotropic radiator. Antenna polarization was circular. Manual and semiautomatic antenna control was used. The radiation power of the ground transmitter could be either 1 or 0.25 kW.

The duration of communication was about 4 min per orbit (T_c = /287
1 h 41.6 m) of the satellite. Relaying of one telephone or 7 teletype channels was accomplished. The satellite was in operation for 34 days and burned up on entry into the atmosphere on Jan. 21, 1959.

Courier IB. The second active radio relay station, Courier, was launched on Oct. 4, 1960 from Cape Canaveral by a Thor-Able-Star booster into an inclined orbit with $i = 28^o$, h_a = 1,270 km, h_p = 970 km and T_c = 106.9 min (ref. 103). The satellite was a plastic sphere 129.5 cm in diameter. The equipment consisted of four transceivers (frequency of 150 Mcps for transmission of commands and 1,900 Mcps for communications), magnetic memory devices and power sources--solar cells and chemical batteries. The total weight of the satellite was 226.7 kg, and the equipment weight was 135 kg.

The equipment operating at the 150 Mcps frequency was analogous to the Score equipment. The passband of the FM receiver at the 1,900 Mcps frequency was 550 kcps, the noise factor was 14 db. Slot antennas with 4 db gain and circular polarization were used for transmission and reception.

The power radiated by the onboard transmitter (vacuum tube) varied in the range of 5-8 W, the power consumed was 80 W. Transmission was FM with deviation of 100 kcps. The frequency stability was $\nu_{inst} = 5 \cdot 10^{-6}$ (ref. 104).

The primary power source consisted of 19,152 silicon solar cells located on the outer surface of the satellite and having an efficiency of 8.5 percent. The buffer stage was 28-32 V 10 A-hr nickel-cadmium batteries.

Communications were established between two ground stations--one at Fort Monmouth, New Jersey, and the other in Puerto Rico--in both delayed relay and real time modes (ref. 104). The ground stations were equipped with dish antennas 8.53 m in diameter with a gain of 19 db in the search regime and /288 43.5 db in the tracking mode and receivers with a noise coefficient of 12 db. Antenna control was semiautomatic. The tracking error amounted to 0.5-0.25°. The beam width of the communications antenna was 18° in the search mode and 1.3° in the tracking mode. The duration of communications was 5 min per orbit of the satellite. Relaying of telephonic and telegraphic radio signals was performed. In order to provide reliable reception of the satellite signals, four receivers were installed at the ground station, using a 4-fold frequency and polarization diversity scheme (refs. 105 and 106). The satellite was in service for one year.

The Telstar System. The development of the Telstar communications satellites began at the Bell System firm in 1958. An experimental communications system has now been constructed. The Telstar system includes five ground communications stations. Two of them are in the U.S.--the first (primary) in Andover, Maine, the second in Holmdale, New Jersey--and three in Europe: the British station in Goonhilly, the French in Pleumeur-Bodou, and the Italian in Fucino. The telemetry data from the Telstar satellites are received by a network of Minitrack stations on the frequency of 136 Mcps.

The first active Telstar relay station (Telstar-1) was launched on July 10, 1962, into an inclined ($i = 44.8°$) elliptic orbit with $h_a = 5{,}600$ km, $h_p = 950$ km, $T_c = 2$ h 38 m, using a Delta booster from Cape Canaveral (ref. 103). The Telstar satellite is a sphere 87 cm in diameter and weighing 77 kg and is intended for active relaying of radio signals in real time. Simultaneous relay of 600 simplex telephone channels, or 12 duplex telephone channels, or 1 television channel is possible. FM operation is used in all cases. The communications frequencies are 6,389.58 Mcps on the uplink and 4,169.72 Mcps on the downlink (ref. 107).

We shall present some data on the various components of the radio channel.

The ground station transmitter consists of four parts: the FM /289 deviator, the modulator-amplifier, the carrier source and the power amplifier. The power amplifier uses a TWT and has a maximal output power of 2 kW. The maximal frequency deviation is $\Delta f_m = 10$ Mcps.

The ground receiving-transmitting antenna at Andover is a horn-reflector antenna with an area of 334 m^2, covered with an inflatáble protective dome. The antenna gain at the transmitting frequency (6,330 Mcps) is 61 db and at the reception frequency (4,170 Mcps) is 58 db. The half power point width of the main lobe is, respectively, 0.16 and 0.23°. The level of the side lobes is less than -20 db (ref. 108).

The maser receiver and power stage of the transmitter are mounted near the horn apex to reduce feeder losses to a minimum (to 1 db).

The Telstar AES relay is designed using a circuit without demodulation of the signal with primary amplification at the IF. The level of the FM radio signal arriving at the relay station from the ground station is equal to about 60 db mW. A crystal mixer is installed at the input to reduce the frequency from 6,389.8 to 90 Mcps.

The conversion losses are 7 db. The IF amplifier with a passband width of 50 Mcps gives an amplification of 65 db in 14 transistorized stages. The noise factor of the onboard receiver is 16.5 db. In the balanced converter consisting of a heterodyne (4,079.73 Mcps) and a shift mixer, the 90 Mcps frequency is increased to 4,169.72 Mcps and applied to a TWT having a gain of 37.5 db. With accounting for the losses in the feeders, the AES antenna radiates a power of 2.1 W.

The onboard receiving and transmitting antennas are an assembly of individual radiators, built in the form of rectangular segments of a waveguide. The waveguide segments form two strips located on both sides of the equatorial plane of the satellite (ref. 108). The transmitting antenna contains 48 segments and the receiving 72. Each radiating element gives a circularly polarized wave. In the equatorial plane both antennas have patterns differing from the pattern of an isotropic radiator by no more than 2 db. /290
In the polar plane the pattern has considerable scalloping. The gain of the transmitting and receiving antennas is equal to 0 db. The FM signal radiated at the frequency of 4,169.72 Mcps by the satellite transmitting antenna is received by the ground station antenna. As an example let us consider the satellite-Andover station link. As we mentioned above, at the frequency of 4,170 Mcps the receiving-transmitting horn antenna has a gain of 58 db. The signal incident on the ground antenna is very weak--its level is equal to about -93 db mW. Directly on the antenna is mounted a ruby traveling wave maser (cooled to 2°K), having a passband of 25 Mcps and a gain of 35 db (ref. 108). The overall noise temperature of the maser is 3.5°K.

The overall noise temperature, referred to the receiver input, at the frequency of $f = 6,150$ Mcps with an elevation angle of 7.5° is equal to 42°K (atmospheric, Earth, protective dome, feeder, maser, following stage noise). Calculations show that the minimal s/n ratio (with the greatest distance of the satellite from the Andover station) at the receiver input on the ground is equal to 15 db. Considering that on the ground use is made of an FM receiver with tracking filter, we can conclude that the input s/n ratio is considerably above the threshold. Thus, we would expect good quality of the signals being relayed. Experience has confirmed this.

The design of the Telstar satellite was intended to provide a service life of two years. The problem of providing power for the onboard equipment over such a long period of time was resolved by the use of solar batteries and storage batteries. The solar batteries are the primary source of the electric power. Elements of the n-on-p type were selected for the cells of the solar batteries. For protection from radiation the elements were covered with a layer of synthetic sapphire 0.76 mm thick. The 3,600 cells were in 50 groups of 72 series-connected cells in each group (ref. 107). The initial /291 nominal power at 28 V was 14 W with any orientation of the satellite relative to the Sun. With the satellite in the shaded zone the equipment is supplied by a battery of 19 nickel-cadmium hermetic storage batteries with an output of 16 V. For supply of the TWT high voltage circuits the 16 V output is converted in a transistorized invertor, then rectified in four rectifiers.

The duration of the communications on the U.S.-Europe link via the Telstar-1 satellite was about two hours a day. The quality of the transmitted television images varied from good to excellent (ref. 108). After more than four months of successful operation the command link failed. It has been established that the most probable cause of the failure is the surface damage to certain transistors of the decoder of the command link as a result of the action of radiation, as the satellite passed through the inner radiation belt. (ref. 107).

On May 7, 1963, the Telstar-2 communications satellite was launched, an 87 cm diameter sphere (just as the Telstar-1), weighing 79 kg. The orbit parameters were: inclination 43°, apogee 12,100 km, perigee 1,060 km, period 221 min.

The parameters of the onboard equipment were the same as those of the Telstar-1. Just as Telstar-1, Telstar-2 was intended for real-time relay of either 600 simplex telephone channels, 12 duplex telephone channels or one television channel. Thanks to the higher orbit than Telstar-1, the time of simultaneous radiovisibility of Telstar-2 in the U.S. and Europe was greater than for Telstar-1. All components of the equipment of the Telstar-2 satellite were subjected to more intensive testing and more careful selection than had been done in the case of the Telstar-1 satellite. Plans have been made for certain modifications in the ground equipment at Andover. Tracking of the satellite by the horn antenna will be accomplished directly from the signals of the microwave beacon. It is also proposed to increase the passband width of the maser of the ground receiver to 50 Mcps.

The Relay Satellite. On Dec. 13, 1962 the Relay communications /292 satellite was launched. The orbit parameters were: inclination $i = 52°$, apogee $h_a = 7{,}420$ km, perigee $h_p = 1{,}319$ km, period $T_c = 3$ h 06 m (ref. 103).

The satellite of the Relay series is an eight-sided prism with maximal dimensions of 74 cm, height of 48.3 cm; the upper portion is a truncated eight-sided pyramid with height of 40.6 cm. The satellite weight is 68 kg (ref. 103). The purpose is the relaying of a single television transmission or 600 simplex telephone transmissions, and also the provision of duplex and other modes of communication. On board the satellite there are two independent repeaters operating with a common antenna, each of which can be energized on command from Earth. The operating frequency for the uplink is 1,725 Mcps, and for the downlink it is 4,170 Mcps (just as for the Telstar satellite) (ref. 109).

At the ground stations (just as with the Telstar satellite system) use is made of 10 kW klystrons having a passband of 10 Mcps. Thus, the width of the signal spectrum (FM) on the uplink cannot exceed 10 Mcps, and with the relaying of television in this case the advantages of FM, which show up with high values of the modulation index, cannot be realized. For the realization of these advantages on the downlink, provision is made on board for frequency tripling (increase of the modulation index by a factor of three) by means of varactors, which provide the required s/n ratio on the downlink. Experiments have shown that with an output power of the onboard transmitter of 10 W and the greatest separation of the satellite and the Andover station, with the relaying of television there is a margin of 6 db above the threshold value (ref. 109).

The satellite receiver is fully transistorized, the heterodynes have quartz crystal stabilization and the frequency multipliers use varactors. A TWT is used as an amplifier.

The onboard receiver has a noise coefficient of 12 db with a passband width of 25 Mcps. The satellite antenna is an omnidirectional slot with circular polarization (for transmission and reception). /293

The primary source of electric energy consists of 8,215 solar cells. The power derived from the solar battery panel is 68 W. When the satellite is in the shade, the equipment power is supplied by silver-cadmium storage batteries. The battery capacity is 100 W at 24 V.

The solar cells are protected from various external influences by a quartz coating (ref. 110). During the lifetime of the first Relay AES, 2,000 experiments were conducted on the establishment of transoceanic communications (color television, telephony, telegraphy, phototelegraphy). At the present time these experiments are being continued with other satellites of the Relay series--Relay-2 and Relay-3.

The Syncom System. On Feb. 14, 1963 a Delta booster launched the first synchronous satellite, Syncom-1, from the Atlantic Test Range. The actual orbit parameters were: inclination 33°51', apogee h_a = 37,022 km, perigee h_p = 34,185 km, period 1426.6 min, drift rate to the east 2.86° per day.

The satellite was manufactured by the Hughes Aircraft firm and had a cylindrical shape with diameter of 71.1 cm, length of 39.3 cm, weight 38.6 kg (ref. 103). The satellite was spin-stabilized (refs. 111 and 112).

The operating frequency of the uplink for the Syncom system was 7,360 Mcps and for the downlink was 1,820 Mcps.

The receiving antenna on the satellite was a slotted dipole with gain of 2 db, the transmitting antenna was an array with gain of 6 db. The noise coefficient of the onboard receiver was 10 db, the passband was 0.5 Mcps. The primary amplification was performed at the IF of 30 Mcps. The HF power amplifier was a TWT with output power of 2 W. The repeater weighed 32 kg and the power consumption was 16.3 W (ref. 113).

The primary power source on board the satellite were 3,840 solar cells with a total power of 28 W at 27.5 V. /294

Communication with Syncom-1 lasted only 20,077 sec, after which observations were made using astronomical methods (ref. 112).

On June 26, 1963 the second synchronous satellite Syncom-2 was launched. The final value of its orbital period was 1,436 min, i.e., closer to the design value than Syncom-1 (ref. 112). The communication equipment on board Syncom-1 and Syncom-2 was the same.

On July 31, 1963 the first tests of the intercontinental telephone communication link between the U.S. and Africa were made. Telephone transmissions, facsimile and one-way teletype transmission were relayed.

The development of an improved Advanced Syncom satellite is being carried forward. It will be cylindrical in shape with a diameter of 147.3 cm and length of 127 cm. Its weight at launch will be 680 kg, and after injection into a stationary orbit the weight will be reduced to 344.7 kg.

The two independent systems for the correction of the location of the satellite in space are designed for a three-year lifetime (ref. 112).

It is planned to install four separate transceivers on board the satellite. A TWT with output power of 4 W will be used as the power amplifier. The receiving antenna will be a colinear array with 8 db gain, and the transmitting antenna will be a phased array with 18 db gain (ref. 104). The uplink will operate at frequencies of 6,020-6,300 Mcps; the downlink will operate at frequencies of 3,990-4,180 Mcps.

The power supply will be a battery of n-p solar cells with capacity of 147 W (ref. 112).

The handling capacity of the communication line of each of the onboard repeaters will be 600 duplex telephone channels or four television channels.

It is planned that the frequency stability will be no less than $\pm 1 \cdot 10^{-7}$. The ground antenna will have a diameter of 25.5 m, and the equivalent noise temperature at the receiver input will be no more than 80°K.

Echo-1. The first of the successfully launched AES intended for /295 passive relay of signals was Echo-1, launched into orbit on Aug. 12, 1960. The satellite weighed 60 kg. This satellite was fabricated in the form of a 30 m diameter inflatable sphere with a shell of mylar plastic covered with aluminum foil with a reflection coefficient of about 98 percent. The orbital parameters were: apogee 1,911 km and perigee 922 km, period of revolution of the satellite around Earth of 118.3 min, inclination 47.2°. The first relaying of telephone signals was accomplished between the JPL station at Goldstone, California and the Bell System Laboratory in Holmdale, New Jersey.

The radio center at Goldstone is equipped with two 25.5 m diameter antennas, one of which is for transmission and the other for reception of signals (ref. 114).

At the radio center in Holmdale a 20 m parabolic antenna is used for transmission and a horn antenna of new design with aperture area of 36 m^2 for reception. The transmission from Goldstone was on 2,390 Mcps, and from Holmdale on 960 Mcps. Circular polarization of the radio signals was used in order to verify the influence of the change of the shape of the passive reflector on the polarization of the received signals.

A series of experiments with the use of the Echo-1 satellite as a passive repeater was conducted by the Collins firm. Communications were established in the north-south direction between stations at Cedar Rapids, Iowa and Dallas, Texas. At the Cedar Rapids station 8.4 m diameter parabolic antennas were used, in Dallas the same antennas were used for transmission, but reception was on a 12 m parabolic antenna (ref. 114).

The transmitters used on the experimental communication line radiated signals with 10 kW of power. The receiving equipment used diode parametric amplifiers with a noise coefficient of 1.5-2 db. The transmission from Cedar Rapids was on 955 Mcps and from Dallas on 810 Mcps.

The following studies were conducted: possibility of SSB /296 transmission with frequency and phase modulation; amplitude and phase distortions; refraction and absorption in the atmosphere, and verification of the reception noise levels and verification of the ground-based equipment.

To facilitate the tracking of the satellite by the ground stations, two small radio beacons were installed in the satellite, radiating a 5 mW signal at a frequency of 107.9 Mcps.

During the course of the experiment it was found that the pressure of the solar rays have a major influence on the orbit of the passive satellite. Thus, under the action of the pressure of solar rays Echo-1 comes 5.7 km closer to Earth in a day (ref. 115). As a result of the fact that there were gas leaks, the shell of Echo-1 deformed and the reflection coefficient diminished. Work is going on presently on the design of more rigid constructions of the metallized spherical reflectors for the passive relay satellites.

In view of the fact that during the reflection from the spherical surface the energy is scattered in all directions, plans are being made for the development in the future of reflectors with complex geometric shapes of the surface (ref. 125).

One of the designs considers the use of a spherical surface divided into a large number of equal segments, metallized in a checkerboard pattern so that opposite each nonmetallized segment on one side of the sphere there is a corresponding metallized segment on the other side.

The overall amount of the reflected energy is markedly increased, since the energy incident on the nonmetallized segments passes through and is reflected from the concave metallized segments of the inner surface of the sphere.

There is a design for the construction of a passive reflector in the form of an icosahexahedron. A reflector of this design acts analogously to a corner reflector, directing the incident energy back along the initial path. Such a reflector with a diameter of 30 m at a frequency of 3 Gcps can provide a directivity coefficient of 52 db.

Echo-2. The second Echo-2 passive satellite was launched on /297 Jan. 25, 1964 from Vandenberg AFB, California, using a Thor-Agena booster into a polar orbit with apogee h_a = 1,300 km and perigee h_p = 1,030 km, T_c = 108.7 min. The satellite weighs 200 kg. The satellite was ejected from a container of dimensions 75 x 100 cm, and after filling with gas the shell diameter of the satellite was more than 40 m (ref. 116).

The USSR (Zimenkakh observatory in Gorkiy), the U.S. and England (Jodrell Bank Observatory, Cambridge) participated in the experiments on the relaying of signals via Echo-2. Plans have been made for experiments on the transmission of ordinary and slowed speech, phototelegraph and teletype communications (ref. 117).

As examples of the solution of the problem on the provision of communication with remote space vehicles let us consider the TRAC(E) system for communication and tracking of space vehicles (ref. 118) and the Telebit system (ref. 119).

The TRAC(E) (Tracking and Communication, Extraterrestrial) system for extraterrestrial communication and tracking provides for tracking and simultaneously permits maintaining radio communication with satellites of Earth, the Moon and with other space vehicles. The system has been under development since 1953 at Caltech JPL, and has found technical embodiment in the Microlock equipment based on the use of the phase method, in which use is made of receivers with passband less than the Doppler frequency shift.

The use of an oscillator with controllable frequency (synchronous heterodyne) makes it possible to use narrow-band amplification prior to detection. Phase autotuning of the frequency in combination with correlation detection provides for simultaneous separation of the telemetry subcarrier frequencies and the Doppler shift, after which the telemetry information is taken from the output of the FM discriminator. Thus the subcarriers are frequency modulated and the carrier is phase modulated, i.e., use /298 is made of FM-PM modulation.

The Microlock system was used for tracking and communication with the Pioneer-III and Pioneer-IV rockets.

The onboard transmitter of the Pioneer-IV space rocket was located in a container compartment with a diameter of 15 cm. Individual components of the circuits and assemblies with semiconductor elements were mounted using printed circuits on micarta disks with an intermediate layer of a thermally insulating material. The vacuum tube stages were mounted directly in the center of the chassis, so that the heating of the container by the tubes was considerably reduced. The total weight of the chassis and the transmitter did not exceed 500 g.

The parameters of the onboard transmitter are presented in table IV-1.

TABLE IV-1

Parameter	Value
Carrier frequency	960 Mcps
Output power	180 mW
Quartz oscillator frequency	40 Mcps
Long-period frequency stability	10^{-6}
Output power stability	$\pm$ 0.5 db
Max. value of random phase modulation	2^{o}
Total weight	360 g
Power consumption	2.4 W
Total volume	458 cm^3
Continuous operation time with onboard battery weighing 2.6 kg	120 hrs
Eff.	7.5 percent

All transmitter stages with the exception of the vacuum-tube power amplifier were transistorized. The onboard transmitter included a transistorized oscillator with quartz crystal stabilization. The oscillator frequency is 40.0021 Mcps. The signal taken from its output is phase modulated (maximal swing one radian) by three telemetry subcarriers, after which it is frequency multiplied by a factor of 24 times, amplified and applied to the onboard antenna. The onboard antenna is practically nondirectional; its gain is only 2.5 db. /299

The primary element of the TRAC(E) system is the 25 m diameter parabolic ground antenna developed for radio astronomical observations and located at Goldstone. With operation at 960 Mcps the width of the main lobe (to the half power points) is 1^{o}, and the gain is 41.8 db (ref. 120).

The receiver noise coefficient in 1,958 was 7.5 db and the overall noise temperature referred to the receiver input was 2,000°K. By the use of low-noise amplifiers (masers) it was planned to reduce the noise temperature to 40°K by 1962.

The receiver passband can be taken equal to 20 or 60 cps. In the radio tracking of the Pioneer-IV the receiver passband was 20 cps (the frequency shift as the result of the Doppler effect in the course of the first minutes of flight was 24 kcps). The receiver threshold level was -154.1 db·mW. With the rocket at a distance of 400,000 km, the level of the received high-frequency signal was -141.1 db·mW, which is 13 db above the threshold level. The tracking station at Goldstone received the rocket signal clear out to a distance of 650,000 km from Earth. Further communication with the rocket was terminated by discharge of the battery supplying the radio transmitter.

An improvement of the Microlock system was the Telebit digital system for space communication (ref. 119). The changeover from the analog system to the digital system was the result of the following factors:

(1) significant advantages in noise immunity of the digital systems over the analog;

(2) convenience of accumulation and storage of digital information on board the satellite and on the ground, and also the convenience of its transmission over the ground communication channels from the point of reception to the processing station.

In the Telebit system use is made of the most advanced of the existing methods of modulation--PCM-RPT. In order to simplify the apparatus, the relative phase keying was accomplished on the subcarriers, which somewhat decreased the effectiveness of the system (by 3-4 db). This solution made it possible to use standard receivers with phase-synchronized loop for /300 tracking the carrier frequency, and required the development of new equipment only for the demodulation of the subcarriers. For the generation of the subcarriers use was made of a tuning fork at a frequency of 1,024 cps, which was also used as the primary synchronizer for the digital equipment. In the development of the block diagrams and the construction of the prototype, particular attention was devoted to the construction of the system with the use of a small number of types of standard elements. The first version of the Telebit system was assembled in four months. It weighed 5.5 kg and required 2.4 W of power. Eight months after the initiation of work, the

Explorer satellite was launched from Cape Canaveral with the first installation of the Telebit system. Somewhat later, on Mar. 11, 1960, the second prototype of the Telebit system, weighing 4.5 kg, was launched on board the satellite Pioneer-V. Both prototypes continued to operate until the radio equipment failed due to failure in the power supply system. These two units operated continuously in the space environment for six months. The unit installed on board the Pioneer-V satellite (transmitter power 5 W) continued to transmit to Earth the required scientific information to a distance of 25 million km (ref. 119).

This summary of the communication systems for the space vehicles launched in the USSR and the U.S., and also the systems for ground communication by means of relaying via AES, gives some idea of the present status of these systems. We see that considerable success has been achieved, both in the resolution of the questions of long range space communication and in the resolution of the questions of the use of AES for surface communication. However, the technical solutions are still far from optimal, and a lot of work remains to be done in the future in both directions.

The mastery of cosmic space, begun in the USSR with the launch of the first artificial satellite of Earth in the history of mankind on Oct. 4, 1957, continued by launchings of spacecraft to the other planets and the unparalleled flights of the pilot-cosmonauts Gagarin, Titov, Nikolayev and Popovich, Bykovskiy and Nikolayeva-Tereshkova, has shown both the complete /301 feasibility of provision for the required forms of communication with the spacecraft, as they travel tens and hundreds of millions of kilometers from Earth, and the possibility of using AES for the relaying of radio signals. The latter permits the development of global and local communication systems in the UHF band without ionospheric disturbances, with a tremendous volume of the transmitted information and a high degree of reliability of transmission.

The rapid development of rocketry and the technology of space communication gives a basis for hope that the communications systems with relaying of information via AES will find wide application in the very near future.

When the manuscript for the present book was already set, the news was heard around the world of the unprecedented group flight of the Soviet cosmonauts on board the multiplace piloted spacecraft Voskhod.

The power of the booster rocket used to inject this space laboratory into Earth satellite orbit with three scientist-cosmonauts and complex equipment exceeded by far the power of the rockets previously used. After a day-long flight and performance of a complex program of investigations, the Voskhod spacecraft landed safely in the intended region of the Soviet Union.

Voskhod is a prototype of those spacecraft in which teams of Earth people will fly to the other planets, to other worlds for the discovery of the secrets of the Universe.

The launch of the Voskhod spacecraft is a triumph of mankind in the conquest of the cosmos, a triumph of Soviet science and engineering, a triumph of the Soviet government.

What we have said here once again confirms the conclusions drawn above of the rapid advances of the spacecraft and the inevitable increase of the flow of information in the radio links for communication with these craft; this in its turn requires additional improvement and optimization of the communications facilities used on board these spacecraft.

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